TM 9-5000-18

DEPARTMENT OF THE ARMY TECHNICAL MANUAL

NIKE I SYSTEMS
TTR TRANSMITTER AND
RECEIVER CIRCUITRY (U)

DEPARTMENT OF THE ARMY • *MAY 1956*

213- CB-MH-20

2 of 2

TECHNICAL MANUAL

NIKE I SYSTEMS

TTR TRANSMITTER AND RECEIVER CIRCUITRY   (U)

TM 9-5000-18}
CHANGES No. 1}

DEPARTMENT OF THE ARMY
WASHINGTON 25, D. C., *26 December 1956*

TM 9-5000-18, 18 May 1956, is changed as follows:

1. The following pen and ink changes are to be inserted:

Page	Paragraph	Line	Changes
2	4a(1)	9	Change period after circuits to semicolon and add "thyratron current pulse triggers the range system in systems with the range modification effected."
7	5a(1)	1	Change "10-volt" to read **40-volt.**
15	Fig. 4	_ _ _ _ _ _ _ _ _	Change "V38" to read **V3B.**
19	8e(2)	4	Delete last sentence. Substitute "this precludes the necessity for high-voltage insulation of the filament transformer."
19	8e(5)	5–8	Delete entire sentence starting with "Bifilar winding" and substitute "Bifilar winding precludes the necessity of high-voltage insulation of the filament transformer."
30	12b	Next to last line	Delete word "inserted."
50	20c(2)	1	Change "arm 1" to read **arm 2.**
50	20c(2)	3	Change "arm 2" to read **arm 1.**
50	20c(3)	1	Change "arm 1" to read **arm 2.**
50	20c(3)	3	Change "arm 2" to read **arm 1.**
71	24b(4)(b)	1	Change "position" to read **portion.**
77	26b(2)	1	Delete first sentence of paragraph.
78	26b(2)	1	Change sentence to read **The 60-mc main surge is the 60-mc AFC signal and is amplified by V3.**
105	32b(1)(a)	15	Change "leads" to **lags.**
105	32b(1)(a)	21	Change "leads" to **lags.**
106	32b(1)(a)	2	Add "due to the limits of Xc of C3" at end of sentence.
123	34b(4)(b)	2, 3, & 4	Change to read **R38, V5B, resistor R45, capacitor C15, resistor R22, V3, through the 250-volt power supply to ground, then from ground to the ground side of capacitor C24.**

2. Remove pages 103 and 104.   Insert revised page 103 and new page 104.

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[AG 413.44 (24 Oct 56)]

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[AG 413.44 (14 May 56)]

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CHAPTER 1

INTRODUCTION

1. PURPOSE AND SCOPE

a. Purpose. The purpose of this text is to provide the reader with an over-all knowledge of the target-tracking radar and with a detailed knowledge of the target-tracking radar synchronizer, r-f and antenna system, transmitter, and receiving systems.

b. Scope. This text presents an over-all block diagram discussion of the target-tracking radar, a detailed block diagram of the target-tracking radar synchronizing, r-f and antenna, transmitting, and receiving systems and a detailed circuit discussion of each. The theory of monopulse radar and operation of the hybrid tee junctions are also discussed in this special text.

2. REFERENCES

References within the body of this text will be of two types: those of the type "6-21.1" refer to TM 9-5000-25 by section (6) and figure (21.1); those of the type "figure 7" refer to figures within the body of the text.

3. DIVISION INTO SYSTEMS

a. This special text contains a detailed functional description and a detailed circuit analysis of the target-tracking radar synchronizing system, target-tracking radar transmitting, r-f and antenna, and receiving systems. This text also covers the introduction to monopulse radar, and the operation of the hybrid (magic) tee. Functionally, the target-tracking radar is divided into eight major systems. Each system carries out one or more of the major functions that contribute to the over-all operation of the radar of the guidance and control equipment. Figure 1-2.1 is a functional block diagram of the target-tracking radar. Each of the systems shown on the block diagram consists of a combination of various components of the radar. The eight major systems are:

(1) Synchronizing system.

(2) Transmitting system.

(3) R-F and antenna system.

(4) Receiving system.

    (5)   Ranging system.

    (6)   Antenna positioning system.

    (7)   Presentation system.

    (8)   Testing, monitoring, and calibrating system.

    b.   As previously stated, this text will cover only the target-tracking radar synchronizing system, transmitting system, r-f and antenna system, and the receiving system, with the aid of functional block diagrams, waveform analysis, simplified schematics, and pictorial diagrams where appropriate. Simplified schematics are not supplied for conventional circuits. In these cases, reference is made to appropriate schematics in TM 9-5000-25.

4. SYSTEM FUNCTIONING AND INTERRELATIONSHIP

    a.   <u>Synchronizing system</u>.

    (1)   The synchronizing system synchronizes the timing of the transmitter pulse with the various range determining and sweep circuits in the target-tracking radar. This system determines the repetition rate (prf) of the radar. To do this, the synchronizing system generates two timing pulses; the sync pulse and the preknock pulse. The sync pulse is applied to the transmitting system and determines the triggering time of the magnetron. The preknock pulse is applied to the presentation system for triggering the sweep circuits, and to the target ranging system for triggering the range determination circuits;

Thyratron current pulses Trigger the range system in system, with the range and modification effected [handwritten annotation]

    (2)   The target-tracking radar is used in conjuction with the acquisition radar for target designation and acquisition. For this reason, the radars must be synchronized. This is done by triggering the synchronizing system of the target-tracking radar with the acquisition preknock pulse from the acquisition radar.

    b.   <u>Transmitting system</u>. The transmitting system generates high-power r-f pulses at a frequency variable between 8,500 and 9,600 megacycles. The sync pulse from the synchronizing system is applied as the input to the trigger generator in the transmitting system. The trigger generator output is applied to a modulator that converts the high voltage from a high-voltage d-c power supply into a series of distinct, 0.25-microsecond pulses. One of these pulses is generated each time the synchronizing system furnishes a sync pulse. These modulator pulses are applied to a magnetron which oscillates for the duration

of each pulse; thereby, generating the high-power r-f signal. The magnetron output is fed by waveguide to the r-f and antenna system.

c. R-F and antenna system.

(1) The r-f and antenna system radiates the high-power r-f energy generated in the transmitting system into space and receives the energy that is reflected back to the antenna when the radar beam contacts a target. It is mainly in the r-f and antenna system that this radar differs from conventional radars. A monopulse system is used to obtain target position information. In most conventional radars, lobe switching or conical scanning is used to obtain target position information, and several pulses must be used to determine the target azimuth and elevation. In the monopulse system, target position data is obtained each time a pulse is transmitted, reflected from the target, and received. This system has 2 pairs of radiating horns or feedhorns (see fig 4-1) and compares the amplitude of a single returned pulse as it appears in the 4 feedhorns. By finding the sum of the differences in the signal strength received by the upper pair of horns and the difference in signal strength received by the lower pair, an azimuth error signal is obtained. Similarly, an elevation error signal is obtained by subtracting the sum of the received signal strength of the upper feedhorn pairs from the sum of the received signal strength of the lower feedhorn pair. The azimuth and elevation error signals generated are used by other systems of the radar to keep the antenna on the target.

(2) The r-f energy in the transmitting system is conveyed by waveguide and accessory equipment in the r-f and antenna system to a metal lens antenna, where the energy from the four feedhorns is radiated in a narrow beam according to a predetermined pattern. The energy reflected from a target is received by the same lens antenna, and after being broken into azimuth and elevation error components, and a sum signal in the r-f system, is conveyed to the receiving system. The r-f and antenna system uses an arrangement of ATR (antitransmit-receive) and TR (transmit-receive) tubes to provide transmit-receive switching of the respective pulses. This system also contains an AFC (automatic frequency control) directional coupler which extracts a sample of the transmitted pulse for generating an automatic frequency control (AFC) signal. This AFC signal is applied to the local oscillator as a control signal to keep the frequency of the local oscillator 60 mc above the transmitter frequency. An additional r-f pulse is extracted by a second directional coupler for power monitoring purposes.

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d. Receiving system. The receiving system converts the X-band signals (called sum, azimuth difference, and elevation difference signals) applied from the r-f and antenna system into 60-mc, i-f signals. Using the three i-f signal components, the receiving system generates d-c azimuth and elevation error signals which are subsequently applied to the antenna positioning system. The sum i-f signal is applied to the target ranging system as the target range video signal. In addition, sum, azimuth, and elevation video signals are applied to the presentation system for the scope displays. Gating of only those pulses received from the target being tracked is accomplished by the track receiver gate being applied to the receiving system from the target ranging system. This gating is necessary to reduce and eliminate echoes received from undesired targets in the vicinity.

e. Ranging system (TM 9-5000-19).

(1) The ranging system measures the time delay between the transmission of the r-f pulse and the reception of the target echo. Since the range of the target is proportional to the time interval between the transmission of the r-f pulse and the reception of the target echo, measuring the time interval determines the range. Three means are provided for controlling the range determining circuits: manual, aided, and automatic. Manual and aided ranging are done through manual ranging controls (handwheels) and are used for acquiring a target; and for target tracking when automatic tracking is impossible. Automatic tracking of the target in range provides smooth range data for the computer. The circuits of the ranging system are triggered by the preknock pulse and transmitted pulse applied from the synchronizing system and transmitting system. These pulses indicate to the ranging system the time of transmission of the r-f pulse and, in effect, determine the zero time of the range scale.

(2) Acquisition of a designated target in range is done by the ranging system and the acquisition radar functioning in conjunction with each other. The ranging system furnishes two signals to the acquisition indicators; a track range gate (TRGA) and an acquisition track range mark (QTRMK). The QTRMK is also sent to the target range slew control unit. There, this signal, representing the range of the target-tracking radar, is compared with the range designated by the acquisition radar. The difference between the target-tracking radar range and the designated range is detected by the range slew control unit which causes the ranging system of the target-tracking radar to slew in or out, in range, to the designated range.

(3) To enable the presentation system to generate a short sweep that is centered around the radar range point, the ranging system furnishes a

3-microsecond expansion pulse. The ranging system also supplies the mixed range video and range notch signals for the indicators in the presentation system. For the angle error detector circuits in the receiving system to determine the antenna pointing error of the target being tracked, the ranging system supplies it with a track receiver gate. This signal distinguishes the target video signal from all other video signals.

f. Antenna positioning system (TM 9-5000-20). The function of the antenna positioning system is to correct the antenna pointing error of the target being tracked. As in range tracking, three modes of operation are provided for controlling the antenna position in azimuth and elevation: manual, aided, and automatic. Manual and aided tracking are done through the appropriate hand-wheels and are used for acquiring a target. Automatic tracking of the target in azimuth and elevation provides smooth coordinate data for the computer. In this mode of operation, the antenna is steered in accordance with d-c azimuth and elevation error signals supplied by the receiving system. The antenna positioning system can acquire a target in azimuth in accordance with azimuth data designated by the acquisition radar. The system can also acquire a target in both azimuth and elevation in accordance with remote position data supplied by a fire direction center (FDC).

g. Presentation system (TM 9-5000-19).

(1) The presentation system is the visual link between the target-tracking radar and the tracking console operators. It is made up of three radar indicators and the associated circuits necessary to display visually the information generated by other functional systems of the tracking radar. The three indicators are: range indicator, azimuth indicator, and elevation indicator. Two additional indicators, the precision indicator (PI) and the plan position indicator (PPI), also appear on the target-tracking console with the other indicators. However, these two are not considered part of the target-tracking radar presentation system and are covered in the acquisition radar functional description (TM 9-5000-9). The PPI displays the acquisition video and the target-tracking radar slant range and azimuth with an electronic cross. The acquisition designation information is shown with a range circle and an azimuth mark. The precision indicator or B-scan presentation displays an area around the electronic cross and covers a sector 5,000 yards in range and 30° in azimuth.

(2) The range indicator uses a modified A-scan presentation and displays a 100-yard movable notch in which the operator centers the target. A 500-yard section centered about the notch is expanded for more accurate

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tracking. The 500-yard expanded section is also displayed on the azimuth and elevation modified A-scan presentations. The sweep length on the three indicators in continuously variable from 20,000 to 100,000 yards with the 500-yard expanded section occupying a constant proportion of the total visible sweep length. Elevation and azimuth manual tracking is performed by pip canceling. Coarse and fine dials displaying the coordinate information are also included and appear on the front panels of the appropriate indicators.

(3) The indicators in the presentation system are synchronized in the radar by the target preknock pulse from the synchronizing system. The azimuth and elevation video signals for the indicators are applied from the receiving system, while the mixed range video and notch and the expansion pulse signals are applied from the ranging system.

h. Testing, monitoring, and calibrating system (TM 9-5000-12). The testing, monitoring, and calibrating system is made up of test equipment built into the radar to make sure that the radar operates properly and with maximum efficiency. Included in this system are units whose functions are: to monitor and to adjust the angle error detectors of the receiving system; to monitor crystal currents, i-f main amplifier bias voltage, and i-f signal levels in the receiving system; to calibrate the ranging system; to provide a means for checking the operation of the magnetron and its associated waveguide in regard to power; and to perform other important functions. The functions are discussed in detail in TM 9-5000-12.

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CHAPTER 2

SYNCHRONIZING SYSTEM

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5. GENERAL

The synchronizing system generates timing signals to synchronize the transmitter pulse with the various range indicating and sweep circuits in the target-tracking radar. The synchronizer provides two 40-volt, 2-microsecond positive pulses for this purpose. The first of these pulses is the preknock pulse which is made available to the sweep and range indicating circuits approximately 23.5 microseconds before the transmitter is triggered. The second output pulse is the synchronizing pulse, which triggers the transmitter. The 23.5-microsecond delay between the preknock and the synchronizing pulses is necessary to begin the functions of the sweep and range unit early enough to permit their use in measuring down to zero range, which corresponds to the leading edge of the transmitted pulse. To allow transmission of accurate acquisition data from the acquisition radar, the target-tracking radar synchronizer is synchronized to the acquisition pulse repetition frequency by the acqui- sition preknock pulse. This is done with the INT-AUTO switch S1 in the AUTO position. With switch S1 in the INT position, the target-tracking radar operates independently as a free-running synchronizer. The operation of the preknock pulse in the target ranging and presentation systems is discussed in detail in TM 9-5000-19. The operation of the synchronizer pulse in the transmitting system is discussed in detail in chapter 3 of this text. The synchronizing system of the target-tracking radar is acquisition and target track synchronizer GS-15616. This unit is located in target-tracking console assembly GS-15513, figure I-16.

a. Inputs and outputs.

(1) The acquisition preknock pulse (positive, 40-volt, 2-microsecond pulse) is an input to target a synchronizer utilized in synchronizing the operation of the target-tracking radar with the acquisition radar. This is necessary to accurately transfer targets from the acquisition radar to the target-tracking radar.

(2) The target preknock pulses (positive, 40-volt, 2-microsecond) initiate the action of the timing, ranging and sweep circuits. Due to the 23.5-microsecond delay in the synchronizer and a 0.9-microsecond system delay, this occurs 24.4 microseconds prior to firing the transmitter.

(3) The target preknock pulse (positive, 40-volt, 2-microsecond) is used to trigger the transmitting system and the i-f test unit.

(4) The test pulse (6-volt, 7-microsecond) output, used in the acquisition system to aline the moving target indicating (MTI) system, is not used in the target-tracking radar.

b. Components. The components of the synchronizing system are: amplifier V1, blocking oscillator V2A, pulse amplifier V3A, sync delay network Z1, switch tube V3B, amplifiers V5A and V5B, and synchronizing pulse blocking oscillator V2B.

c. Output distribution (fig 2-1).

(1) The target synchronizing pulse (40-volt, 2-microsecond) is applied to the following:

(a) The target trigger generator, where it triggers a single-swing blocking oscillator.

(b) The i-f test unit, where it triggers circuits that develop an i-f pulse for adjusting the target-tracking radar receiving system.

(2) The target preknock pulse (40-volt, 2-microsecond) is applied to the following:

(a) The range unit assembly, where it triggers circuits that develop the tracking range mark (TRMK), tracking range gate (TRGA), and the acquisition tracking range mark (QTRMK).

(b) The delay amplifier, where it triggers a flip-flop multivibrator and makes possible a comparison of range settings of the acquisition and tracking radars possible. Slewing of the tracking range to the acquisition range is dependent upon this comparison.

(c) Three target-tracking sweep generators, where it initiates sweeps for the target-tracking indicators.

(d) The video error signal unit where it triggers circuits to permit pip canceling on the target-tracking indicators.

(e) The test delay unit, where it triggers circuits that develop an artificial target which may be varied in range to check and adjust various circuits of the target-tracking radar.

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6. BLOCK DIAGRAM DISCUSSION (fig 2-1)

a. **Driver V1.** Tube V1 amplifies the preknock pulse from the acquisition radar when INT-AUTO switch S1 is in the AUTO position. With switch S1 in the INT position, the acquisition preknock pulse is grounded through a resistor and there is no input to the driver. The amplified acquisition preknock pulse from the driver synchronizes the blocking oscillator with the acquisition radar pulse repetition frequency (prf).

b. **Blocking oscillator V2A.** Tube V2A, with INT-AUTO switch S1 in the INT position, is a free-running blocking oscillator. With the switch in the AUTO position, preknock pulses from the acquisition radar are amplified by V1 and drive the blocking oscillator. The acquisition preknock frequency is slightly higher than the free-running frequency of the blocking oscillator. The blocking oscillator locks to the frequency of the preknock pulse, giving an output which is synchronized to the prf of the acquisition radar. The positive output pulse goes to the range and presentation circuits as the track preknock pulse and to the delay section of the synchronizer where the 23.5-microsecond delay between the preknock and the synchronizer pulse is introduced. The preknock pulse is a 40-volt pulse and is about 2 microseconds in duration.

c. **Tripper V3A.** Tube V3A inverts the positive preknock pulse and the negative output pulse triggers monostable multivibrator V4 and starts delay network Z1 in the delay pulse generator.

d. **Multivibrator V4.** Tube V4 is a monostable multivibrator used as a switch to start operation of the Z1 delay network. The multivibrator is triggered by the target preknock pulse. The delay time of the multivibrator is great enough to allow network Z1 to discharge. After delay network Z1 discharges, the multivibrator returns to its quiescent state until triggered by the next preknock pulse.

e. **Delay pulse generator V3B and Z1.** Delay network Z1 is an R-C network in which the discharge time from preknock is less than the period of the multivibrator. For this reason, jitter and variations of time delay, which are inherent in the monostable multivibrator, have no effect on the succeeding circuitry. Network Z1 controls the on-off time of diode V3B and produces a 23.5 microsecond, negative, square pulse at its output. The leading edge of the square pulse coincides in time with the track preknock pulse, and the trailing edge is set by adjustment of the network discharge time.

f. **Delay pulse amplifier V5A.** Tube V5A amplifies the output from V3B and the amplified pulse is applied to the grid of V5B.

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g.   Driver V5B. Stage V5B is the driver for synchronizing pulse blocking oscillator V2B. Pulse transformer T3 in the plate circuit of V5B differentiates the output of V5B so that V2B will be triggered on the trailing edge of the delayed pulse.

h.   Synchronizing pulse blocking oscillator V2B. Tube V2B is triggered by the trailing edge of the differentiated delayed pulse from V5B. The blocking oscillator action provides the 40-volt, 2-microsecond synchronizing pulse which triggers the transmitting system. Because of cable delays of approximately 0.9 microsecond, the transmitter is triggered about 24.4 microseconds after preknock.

7. DETAILED CIRCUIT OPERATION (fig 2-2)

a.   Driver V1. With the INT-AUTO switch S1 in the AUTO position, the acquisition preknock pulse is amplified by V1 and the waveform is differentiated by windings 5-6 of transformer T1 in the plate circuit of V1. With the INT-AUTO switch in the INT position, there is no input to V1, so it is no longer effective in driving the preknock blocking oscillator. Tube V1 is a 6AU6 pentode whose grid bias is determined by resistor R3 and capacitor C2. The effective plate voltage is determined by resistor R4 and is decoupled by capacitor C3A. Resistor R1 is the screen dropping resistor and capacitor C1 is the screen bypass capacitor. The positive spike of the differentiated preknock pulse is transformer coupled to the grid of the blocking oscillator.

b.   Blocking oscillator V2 (figs 2 and 3). Tube V2 is a self-pulsing blocking oscillator designed to give a sharply defined output pulse about 2 microseconds in duration. The period of time between pulses is determined by the time constant of capacitor C4 and the series combination of resistors R11 and R6. Resistor R6 is the adjustment to give the desired period. For normal operation, it is adjusted to give a period slightly greater than the acquisition preknock period. Starting with the leading edge of the output pulse, operation of the blocking oscillator, under free-running conditions, is as follows: plate current flow in V2A starts building up a magnetic field about winding 1-2 of transformer T1. The expanding magnetic field causes a voltage to be induced across winding 3-4 of transformer T1 causing the grid to go positive, resulting in a further increase of plate current. The buildup of the magnetic field continues until plate current saturation, at which time the magnetic field about winding 1-2 ceases to expand. The induced positive voltage applied to the grid of V2A causes grid current flow which charges capacitor C4 (fig 1). At plate current saturation, when the magnetic field about winding 1-2 ceases to expand, there is nothing to sustain the buildup of the magnetic field about winding 3-4. Consequently, the magnetic field built up about winding 3-4 because of plate current flow begins to collapse. This applies a negative voltage to the grid of V2A,
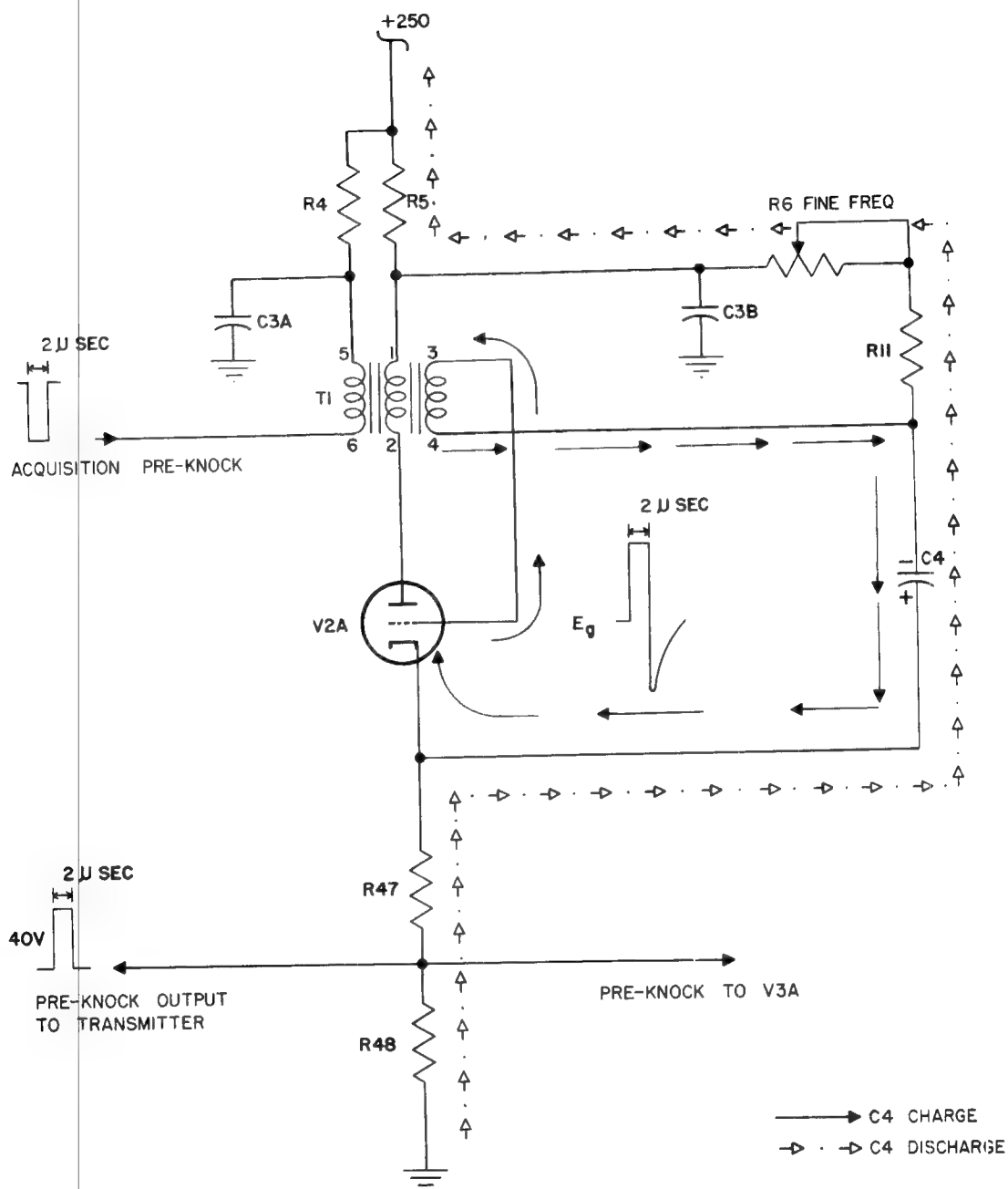
10

Figure 1. Blocking oscillator.

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decreasing plate current. The decrease in plate current results in a collapse of the magnetic field about winding 1-2, which adds to the collapse of the field about winding 3-4, driving the grid even more negative; thus, cutting the tube off abruptly. The grid voltage takes a large negative swing and returns to a value established by the grid current charge on capacitor C4, at which time capacitor C4 begins an exponential discharge through resistors R11 and R6. The amount of time required for capacitor C4 to discharge to a value where V2A can begin conduction determines the period of the blocking oscillator. The large amount of regeneration on both the buildup and collapse of the magnetic field results in a very sharp leading and trailing edge of the output waveform. The output is taken across resistor R48 in the cathode circuit of V2A. As a result, the output is a positive rectangular waveform corresponding in time to the conducting period of V2A. The acquisition preknock pulse frequency is slightly higher than that of V2A. The acquisition preknock pulse, which is differentiated by winding 5-6 of transformer T1, triggers V2A slightly before capacitor C4 discharges, causing V2A to lock on at the same frequency. Thereby, the two radars, acquisition radar and target-tracking radar, synchronize. Resistor R5 is the plate dropping resistor for V2A with capacitor C3B as the plate bypass capacitor. Resistors R47 and R48 are cathode resistors, with the output developed across resistor R48. With capacitor C4 connected to the cathode, there is a minimum voltage excursion between cathode and grid on the positive swing of the grid. The output signal (fig 2) is cathode-coupled to V3A through capacitor C5.
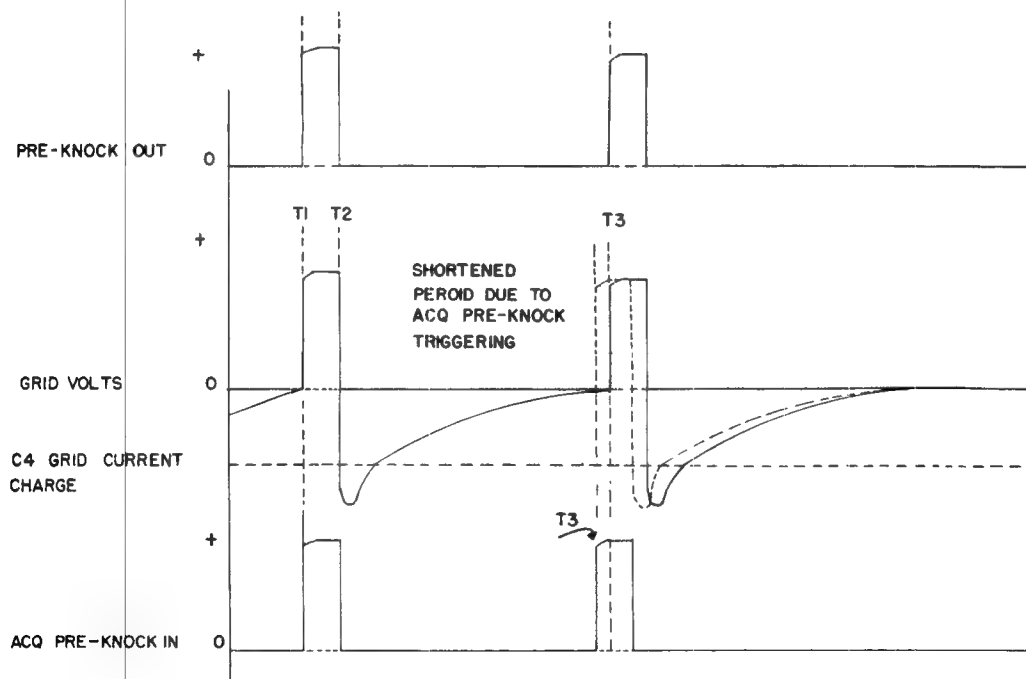

Figure 2. Blocking oscillator waveforms.

c. **Tripper V3A.** The target preknock pulse is differentiated by the action of capacitor C5, resistors R13 and R14 in the grid circuit V3A. Bias for V3A, keeping the tube cut off in the absence of an input pulse, is provided by voltage divider R12-R13. Resistor R14 is also a parasitic suppressor. Resistor R15 is the common plate load for both V3A and V4A. The differentiated preknock pulse appears at the plate of V3A as a negative spike which triggers multivibrator tube V4B and starts the delay circuit operation.

d. **Multivibrator V4 (fig 3).** Tubes V4A and V4B form a monostable multivibrator. In the quiescent state, V4B is conducting, with V4A held cut off by voltage divider R16-R17. The grid of V4B is slightly positive because of the voltage drop resulting from grid current flow through resistor R22. Since V4A is cut off, there is no voltage drop across resistor R15, so the plate voltage is +250 volts. Capacitor C7, which controls the operation of the multivibrator, has +250 volts on its upper plate and is only slightly positive on its bottom plate, resulting in a charge of almost 250 volts. The negative trigger from V3A is coupled through capacitor C7 to the grid of V4B, reducing the tube conduction. Thereby, the plate voltage goes positive. The positive-going plate voltage from V4B is coupled to the grid of V4A through capacitor C6, driving V4A into heavy conduction with a resulting decrease in plate voltage (fig 3). Capacitor C7 must then discharge to allow the grid voltage of V4B to rise to a value positive enough for V4B to start conduction and return the multivibrator to its quiescent state. The discharge time of capacitor C7 is just long enough for the succeeding time delay network to complete its operation before the multivibrator reverts to its quiescent state. (See figure 5 for multivibrator and delay circuit waveforms.)

e. **Delay pulse generator V3B and Z1 (figs 4 and 5).** Network Z1, in conjunction with multivibrator V4 and tube V3B, establishes the 23.5-microsecond delay between the preknock pulse and the synchronizing pulse. During the quiescent period of multivibrator operation, the delay network capacitor is charged to the difference between the +250 volts on the plate of V4A and the small positive voltage on the plate of conducting tube V3B. When the multivibrator is triggered, the plate voltage of V4B drops from +250 volts to about 20 volts. This change in voltage cuts V3B off and V3B is held at cutoff until the delay network capacitor discharges enough for the tube to start conducting again. The discharge time of the network is set to 23.5 microseconds by adjusting SYNC DELAY potentiometer R21. The delay network is temperature controlled by a heating element, which is powered from a 6.3-volt a-c source to prevent drift of the delay time with a resulting unstable radar operation. The output is a negative rectangular pulse with a trailing edge delayed 23.5 microseconds from the preknock pulse. The output is taken across resistor R41 and applied to the grid of V5A.
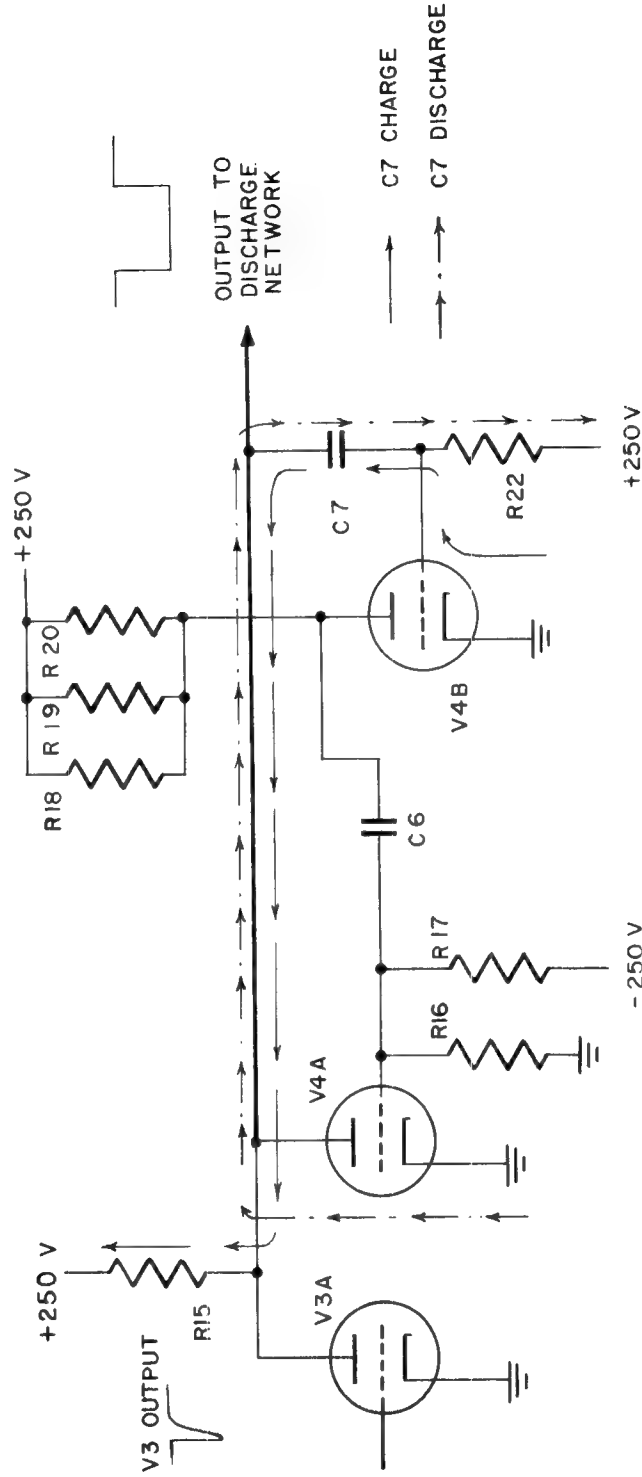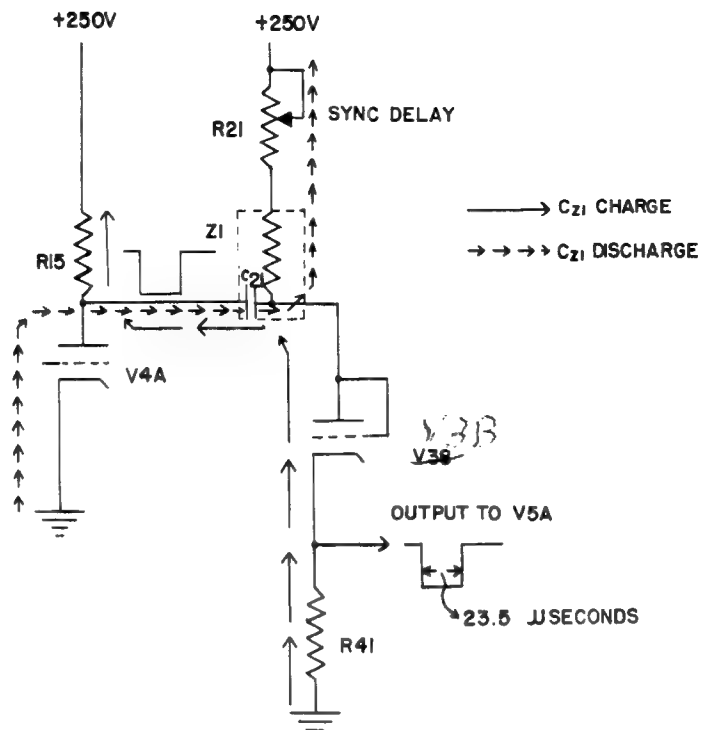
Figure 3. Multivibrator.

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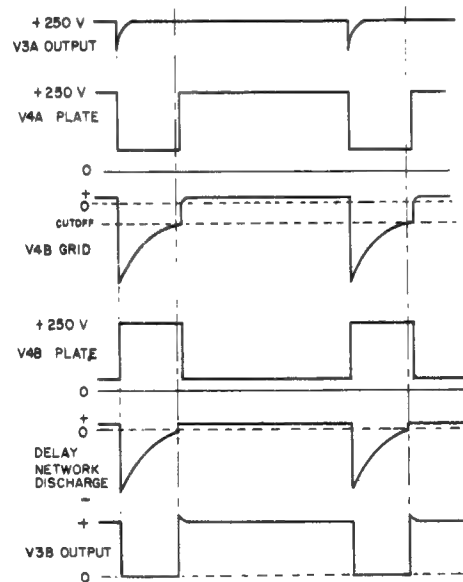Figure 4. Delay pulse generator.

Figure 5. Multivibrator and delay circuit waveforms.

f. Delay pulse amplifier V5A. The delay pulse amplifier inverts and ampli-
fies the 23.5-microsecond rectangular pulse from the delay pulse generator.

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The plate load is the parallel combination of resistors R23 and R24. Bias is furnished by unbypassed cathode resistor R25. The output is a positive 23.5-microsecond pulse which is coupled to the grid of V5B through capacitor C8.

g. <u>Driver V5B.</u> Tube V5B amplifies and inverts the output from V5A. The output from V5B is differentiated by winding 5-6 of transformer T3 to give a negative and positive pulse at times coincident with the leading and trailing edges of the rectangular waveform. Bias for V5B is provided from -250 volts applied across voltage divider R27-R28. The lower half of the voltage divider (R27) is bypassed by capacitor C9. The negative pulse, corresponding to the leading edge of the input waveform, has no effect on blocking oscillator V2B. The positive pulse, corresponding to the trailing edge, triggers the blocking oscillator into operation to generate the synchronizing pulse for the target-tracking radar.

h. <u>Synchronizing pulse blocking oscillator V2B.</u> Tube V2B is a blocking oscillator similar to V2A, except that it is designed as a single-swing blocking oscillator and will not free-run. Bias to prevent self-pulsing is applied by the voltage divider consisting of resistors R32, R33, and cathode resistor R30. Resistor R33 is bypassed by capacitor C11 for decoupling voltage variations from the -250-volt bias voltage supply. Resistor R29 is the plate dropping resistor and capacitor C10 is the plate bypass (decoupling) capacitor. The output pulse is developed across resistor R30 in the cathode circuit. This pulse is a positive 40-volt, 2-microsecond pulse that is applied to the target-tracking radar transmitting system and blocking oscillator V6B. Tube V6B is not used in the target-tracking radar.

CHAPTER 3

TTR TRANSMITTING SYSTEM

8. OVER-ALL SYSTEM BLOCK DIAGRAM

a. General.

(1) The target-tracking radar transmitting system provides the r-f and antenna system with high-energy radio frequency in the form of repetitive pulses. The repetition rate of 1,000 pulses per second is determined by the synchronizing system. The 0.18-microsecond duration of each pulse is determined by the transmitting system itself. Each pulse, in turn, is applied to the r-f and antenna system with a normal peak power of approximately 200 to 400 kilowatts, at a radio frequency which is variable within the range of 8,500 to 9,600 megacycles. The target-tracking radar transmitting system is made up of track trigger generator GS-15460, slipring assembly GS-15536, track modulator GS-15461, a pulse transformer, and the 5780 magnetron with its tuning drive and associated circuitry. Only a portion of the slipring assembly is used in the transmitting system. All components of this system are located on the target antenna trailer GS-15520.

(2) Figure 3-1.1 is a block diagram of this system. The trigger generator receives the 40-volt, 2-microsecond pulse from the synchronizing system at the rate of 1,000 pps, and produces a 230-volt, 4-microsecond pulse, which is applied to the track modulator. The track modulator, with its associated high-voltage power supply, produces from this pulse a 6,500- to 8,500-volt, 0.25-microsecond pulse. This pulse is transferred through sliprings to the pulse transformer, which steps up the pulse to 30 to 35 kilovolts, and applies it to the magnetron. Each pulse causes the magnetron to oscillate at the frequency to which it is tuned within the 8,500- to 9,600-megacycle range. The output of the magnetron is fed by waveguide to the r-f and antenna system.

b. Trigger generator.

(1) Location. The trigger generator is located in the equipment enclosure (roadside) of the target-tracking radar antenna.

(2) Purpose. The trigger generator receives the sync pulse from the target synchronizer and builds it up enough to trigger the modulator.

(3) Inputs. There is one input; a 40-volt, 2-microsecond positive sync pulse from the target synchronizer.

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(4) Outputs. One output is supplied. This is a 230-volt, 3- to 5-microsecond, positive trigger pulse, which is applied to the modulator.

(5) General operation. The sync pulse is applied to a single-swing blocking oscillator circuit which increases the duration time to from 3 to 5 microseconds and amplifies the pulse from 40 volts to 230 volts by doubling the amplitude of the blocking oscillator output through transformer action. The 230-volt, 3- to 5-microsecond pulse is then applied to the modulator.

c. Modulator.

(1) Location. The modulator is located in the antenna equipment enclosure, next to the trigger generator.

(2) Purpose. The modulator generates and shapes the pulse that will cause the magnetron to oscillate and generate an r-f pulse.

(3) Inputs. This chassis has one input; a 230-volt, 3- to 5-microsecond pulse from the trigger generator.

(4) Outputs. The single output is the negative, 6,500- to 8,500-volt, 0.25-microsecond pulse which is applied to the primary winding of the pulse transformer.

(5) General operation. This unit is composed chiefly of a pulse-forming network (Z1), an input thyratron switch tube, a charging diode, and a reverse-current diode. The Z1 network is charged by voltage from the high-voltage power supply and operates as a d-c resonant charging circuit. The charge on the pulse-forming network is trapped by the action of the charging diode. The thyratron switch tube, when triggered, causes the pulse-forming network to discharge and supply the pulse which, after amplification, will fire the magnetron. The reverse-, current diode prevents the pulse-forming network from acquiring a negative charge, thereby permitting it to start charging from the same level at the beginning of each charging cycle. A reverse-current meter in the circuitry will indicate the value of reverse current (average 20 ma) and give an indication of magnetron malfunction by indicating when reverse current has become excessive (over 30 ma).

d. Sliprings.

(1) Location. The sliprings are on the trunnion assembly of the tracking radar antenna.

(2) Purpose. The sliprings provide electrical continuity between the rotating parts of the antenna assembly and the stationary parts.

(3) Inputs. The input from the transmitter is the negative 6,500- to 8,500-volt, 0.25-microsecond pulse from the modulator.

(4) Outputs. The 6,500- to 8,500-volt pulse is conducted through the sliprings and is applied to the pulse transformer.

(5) General operation. The sliprings are a group of 89 inner ring segments. A like number of contactor arm assemblies are mounted mechanically in a manner to maintain circuit contact with the rings at all times when the antenna traverses on its trunnion assembly. These assemblies maintain circuit continuity in a flexible manner. This permits passage of signals through the assembly without changing them.

e. Pulse transformer.

(1) Location. The pulse transformer is in the track r-f unit near the magnetron.

(2) Purpose. The pulse transformer receives the negative modulator pulse and steps it up without phase inversion. The stepped-up high-voltage pulse to the magnetron cathode is applied through bifilar windings. ~~This prevents a great potential difference from existing between cathode and heater and probable heater breakdown.~~ This precludes the necessity for high-voltage insulation of the filament transformer.

(3) Inputs. The pulse transformer input is the negative 6,500- to 8,500-volt, 0.25-microsecond pulse which is applied to the primary winding terminals from the modulator.

(4) Outputs. The output of the pulse transformer is a negative 30- to 35-kilovolt, 0.25-microsecond pulse with a step in the leading and trailing edges. This pulse is applied to the magnetron cathode.

(5) General operation. Transformer T4, the pulse transformer, is wound so that there is a step-up ratio from primary to secondary of 1:4. There is no phase inversion of the voltage output from the secondary as is normally encountered through transformer action. The secondary voltage is applied to the cathode of the magnetron. Bifilar winding averts the danger of a high potential difference existing between the cathode and heater causing possible arc-over and consequent heater breakdown. The bifilar winding design of the transformer is such that it places the same potential on the heater and cathode during the application of high-voltage triggers. When the high-voltage pulse is

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applied to the magnetron cathode, the magnetron will oscillate at the resonant frequency of the slug-tuned cavities.

f. Magnetron.

(1) Location. The magnetron is mounted in the track r-f unit. With the cover removed from the target-tracking radar and the observer facing the lens assembly from the rear of the r-f unit, it is to the upper left of the assembly.

(2) Purpose. The magnetron receives the output of the pulse transformer and oscillates at a frequency determined by the resonant size of its cavities. The size of the cavities is determined by the setting of the tuning slugs controlled by the tuning drive unit. It delivers an r-f output when triggered by the pulse transformer. This radio frequency is necessary to determine target information.

(3) Inputs. The input to the magnetron (other than heater voltage) is the negative 30- to 35-kilovolt, 0.25-microsecond pulse from the pulse transformer.

(4) Outputs. The output of the magnetron is an r-f pulse; 0.18 microsecond in duration; peak power, 200 to 400 kilowatts; frequency range, 8,500 to 9,600 megacycles.

(5) General operation. Application of the pulse from the transformer to the magnetron cathode causes electron flow from the cathode toward the plate (anode). This electron flow will bring about oscillations in the slug-tuned cavities of the magnetron. These oscillations produce an output pulse whose frequency is determined by the cavity size and whose output power is on the order of 200 to 400 kilowatts. The magnetron's frequency range is 8,500 to 9,600 megacycles per second. The duration of this transmitted pulse is 0.18 microsecond.

g. High-voltage power supply.

(1) Location. This unit is located in the upper left corner of the equipment enclosure (roadside).

(2) Purpose. Its supplies a high-potential d-c voltage output to charge the pulse-forming network, Z1, in the modulator.

(3) Inputs. The inputs are 120-volt, 3-phase, a-c voltages from the system a-c distribution.

(4) Outputs. The output is a 6,500- to 8,500-volt d-c voltage applied to the tracking modulator.

(5) General operation. This power supply is a full-wave type rectifier which has a-c inputs. A transformer and two diodes rectify and amplify this voltage into a high d-c potential. It is applied to the modulator for charging the pulse-forming network (Z1). The high-voltage d-c potential is determined by the setting of the high-voltage supply control on the control panel of the target-tracking radar console. This portion of the transmitting system wil not be discussed in this text but is covered in detail in TM 9-5000-17 .

h. Magnetron tuning drive.

(1) Location. This circuitry is located in the track r-f unit of the antenna mount assembly. Controls are located in the r-f unit and on the target-tracking console assembly.

(2) Purpose. The tuning drive assembly positions the tuning slugs in the cavities of the magnetron so as to adjust the physical size of the cavities for resonance at the desired frequency of transmission.

(3) Inputs. The inputs consist of a-c driving and motor excitation voltages.

(4) Outputs. The output of the tuning drive is mechanical slug positioning.

(5) General operation. There are two FREQUENCY INCREASE-DECREASE switches associated with the tuning drive. One is on the target console control panel in the radar control trailer and the other is on equipment panel A of the track r-f unit at the antenna. These switches are spring-loaded to keep them in a midposition (neutral). When either is pressed to the DECREASE position, voltage is applied to the driving motor in such phase that it causes the slug-positioning mechanism to withdraw the slugs and increase the cavity size, which decreases the frequency. If either switch is operated to the INCREASE position, the tuning drive motor will be driven in the opposite direction, resulting in a positioning of the slugs deeper into the cavities and a decrease in cavity size results in an increase of magnetron frequency of oscillation. A FREQUENCY meter, located on the control panel at the target-tracking console, is connected in the tuning drive circuitry in such a manner as to indicate relative frequency of the magnetron tuning.

9. TRIGGER GENERATOR (fig 3-3)

a. General. Track trigger generator GS-15460 is in azimuth drive equipment enclosure GS-15692 on the target-tracking radar antenna trailer. This

unit consists of a sync pulse amplifier V1A, blocking oscillator V1B, and cathode follower V2. The unit receives a positive 40-volt, 2-microsecond pulse (1,000 pps) at input jack J1, through a coaxial cable from acquisition and target track synchronizer GS-15616, and delivers a 230-volt, 4-microsecond pulse through output jack J2 to track modulator GS-15461. Input jack J3 is not used in the target-tracking radar, but is provided for use when this unit is installed in the missile-tracking radar. Jack J3 is connected to winding 1-2 of transformer T3, which is connected to cause phase inversion across the transformer. This phase inversion permits the negative missile-tracking radar coder pulse to trigger the circuit.

b. Block diagram discussion (fig 3-2).

(1) Pulse transformer T3. Input transformer T3 is a 1:1:2 step-up pulse transformer, connected so as to prevent any phase inversion between windings 3-4 and 5-6. Winding 1-2 is connected so as to cause phase inversion when used in conjuction with winding 5-6. This connection, however, is used only when the radar is used as a missile-tracking radar. The pulse transformer has its natural function of impedance matching.

(2) Driver V1A and T1. Tube V1A and winding 1-2 of transformer T1 make up a sync pulse amplifier (driver) which couples the pulse to the grid of blocking oscillator tube V1B, which has a common load with V1A. Transformer T1, like transformer T3, is a 1:1:2 step-up pulse transformer, but is connected so as to cause phase inversion between windings 1-2 and 3-4. Winding 3-4 is in the grid circuit of the blocking oscillator tube, V1B.

(3) Blocking oscillator V1B, T1. The blocking oscillator forms pulses that are of a predetermined width. In the target-tracking radar trigger generator, the input pulse of 2 microseconds is, in effect, stretched to 3 to 5 microseconds. For discussion, the width of the 3- to 5-microsecond pulse will be referred to as a 4-microsecond pulse. Again, pulse transformer T1 is used in its natural role of impedance matching.

(4) Cathode follower V2. The cathode follower is an isolation amplifier that prevents the entrance of the high-voltage pulses produced in the track modulator unit from entering the blocking oscillator.

c. Detailed circuit discussion (fig 3-3).

(1) Sync pulse amplifier. The 1-2-3 (V1A) section of V1 and its associated components make up the sync pulse amplifier (driver). A positive sync

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pulse entering jack J1 appears as a positive pulse on the grid of V1A by the transformer action of transformer T3. This pulse entering jack J1 is the synchronizing pulse. Tube V1A is normally cut off with a fixed -31 volts applied to the grid from voltage divider R4-R6. When the positive pulse from terminal 6 of transformer T3 appears on the grid, tube V1A conducts heavily and produces a large negative pulse on terminal 1 of transformer T1. Resistor R13 is a grid limiting resistor.

(2) Blocking oscillator. The 6-7-9 (V1B) section of V1 and its associated circuit components make up the blocking oscillator. Voltage divider R1-R2 supplies approximately -31 volts to the grid of V1B. This fixed bias voltage normally cuts off the tube. The negative pulse from V1A on terminal 1 of transformer T1 appears as a positive pulse on terminal 4 of transformer T1. Thus, the grid of V1B goes positive, permitting the tube to conduct. When the tube conducts, still more current flows through winding 1-2 of transformer T1, which causes terminal 1 to become more negative. This, in turn, produces a more positive pulse on terminal 4 of transformer T1 and the grid of V1B, making the tube conduct more heavily. When the plate of V1B reaches its maximum conduction of current (saturation), the field produced in transformer T1 starts collapsing. This action causes a very large negative pulse to appear on terminal 4 of transformer T1 which immediately drives the grid bias beyond cutoff. The tube remains cut off (by the -31 volts appearing on the grid) until the next sync pulse is applied to the unit. This blocking oscillator produces a single cycle, and the action described is very rapid. The length of the pulse depends on the leakage inductance and distributive capacitance of transformer T1. The large positive pulse (approximately 500 volts) appearing at terminal 6 transformer T1 is applied to the grid of tube V2.

(3) Cathode follower V2. Cathode follower tube V2 consists of two triode sections connected in parallel. This tube isolates blocking oscillator V1B from the high-voltage pulses produced in the track modulator. Resistors R7, R10, and R12 are parasitic suppressors. They prevent oscillations within the tubes. The large positive pulse appearing at terminal 6 of transformer T1 is applied to grids of V2 through resistors R10 and R12. This pulse causes the tube to draw plate current until the grid potential (and cathode potential) is equal to the plate potential. At this point, plate action by both grids raises the cathode well above plate potential. The action just described produces a positive 230-volt, 4-microsecond pulse across cathode resistor R9. This positive pulse is shunted across capacitor C5, because capacitor C5 does not have good high-frequency response. Capacitor C6 is a mica capacitor and has a

very good high-frequency response. This parallel combination of capacitors C5 and C6 permits good reproduction of the pulse. The output pulse is passed through a coaxial cable to the track modulator. The cathode of V2 is connected to the filament so that the breakdown voltage rating between cathode and filament is not exceeded during the pulse period.

(4) Component usage. Capacitors C1A, C1B, C2, C3, and C7, along with the 2-microfarad capacitor added by field change 367N, provide decoupling. Transformer T2 provides filament voltage for V1 and V2. Resistors R5 and R8 dampen any ringing oscillations that may occur across the pulse transformers. Test point 1 (TP1) is for testing the bias voltage applied to the blocking oscillator. Test point 2 (TP2) is connected directly to the output of the unit and serves as a point at which to monitor the trigger pulse.

10. TRACK MODULATOR (fig 3-5.1)

a. General (fig 3-4.1). Track modulator GS-15461 accepts the trigger pulse from track trigger generator GS-15460, and delivers a 7,500-volt, 0.25-microsecond pulse to the pulse transformer, which fires the magnetron. (The 7,500-volt figure is arbitrary.) The track modulator unit consists of hydrogen thyratron V1, reverse-current diode V2, charging diode V3, pulse-forming network Z1, and associated circuit components. This unit is located in azimuth drive equipment enclosure GS-15692 on the target-tracking radar antenna trailer.

b. Block diagram discussion (fig 3-4.1).

(1) Charging coil L1. Coil L1 is part of the d-c resonant charging circuit and allows the pulse-forming network to attain a charge of nearly twice the high-voltage d-c input of 6,500 to 8,500 volts from the high-voltage power supply. This charging coil is used in conjunction with the charging diode V3 to accomplish its mission.

(2) Charging diode V3. Charging diode V3, like coil L1, is part of the d-c resonant charging circuit and also helps pulse-forming network Z1 to attain the desired charge. In conjunction with charging diode V3 and coil L1, coil L3 is also in the circuit.

(3) Pulse-forming network Z1. The pulse-forming network applies a high current supply to the primary of the pulse transformer in the cathode and heater circuit of the magnetron. This current is applied when thyatron V1 is triggered by the pulse from the trigger generator,

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causing pulse-forming network Z1 to discharge through the primary of the pulse transformer in the magnetron cathode circuit.

(4) Thyratron V1. Thyratron V1 provides the necessary trigger for pulse-forming network Z1 to discharge; thereby, providing the necessary current to couple a high-voltage pulse to the cathode circuit of the pulse transformer in the magnetron. The thyratron (V1) is triggered by the 230-volt, 4-microsecond pulse from the trigger generator.

(5) Reverse-current diode V2. The reverse-current diode is in the circuit to prevent the pulse-forming network from attaining a negative charge when the network discharges. This will allow the pulse-forming network to always have the same level of charge when the new charging cycle is started.

(6) Reverse-current meter. The reverse-current meter measures the amount of reverse current drawn by reverse-current diode V2. With this meter, it can be reasonably determined whether the magnetron is causing a mismatch and to what extent.

c. Detail circuit discussion (fig 3-5.1).

(1) Charge of pulse-forming network. Refer to figure 6 for a simplified schematic of the track modulator. There is +8,500 volts at high-voltage d-c input jack J2, supplied by track high-voltage power supply GS-15465. This is the source of d-c power which charges network Z1. The +8,500 volts, applied through inductance L1 to the plate of charging diode V3, causes V3 to conduct. The charging path of the Z1 network is as follows: electrons are attracted from the upper (+) plates of the capacitors in the Z1 network, through L3, V3, L1, and the power supply, ground and the primary of the pulse transformer through jack J3, to the lower (-) plates of the capacitors in network Z1. This charging process takes place between trigger pulses from the trigger generator which occur at a 1,000-pps rate. Charging diode V3, which passes current in only one direction, clamps the potential across network Z1 at the highest value reached during the charging process. The capacitance of pulse-forming network Z1 and the 6.5-henry inductance of L1 act as a series resonant circuit. Theoretically, this action produces a charge on network Z1 equal to twice the value of the +8,500 volts applied. Since the circuit is not 100 percent efficient, the voltage across the capacitors rises only to about 16,000 volts.

(2) Discharge of pulse-forming network (fig 6). The 230-volt, 4-microsecond pulse from the trigger generator appears at pulse input jack J1.

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This pulse is applied to the control grid of thyratron V1 and causes ionization in the tube. The resulting conduction is heavy and almost forms short circuit between cathode and plate. When V1 conducts, a low resistance path is completed across the plates of capacitors in network Z1. This path is traced from the lower (-) plates of the capacitors, through jack J3, the pulse transformer, ground, the 0.5-ohm resistance of the R2 assembly, V1, part of inductance L3, to the upper (+) plates of the capacitors. It is apparent that tube V1 performs a switching function; between pulses, its nonconduction opens the low resistance path, allowing network Z1 to be charged. During pulses, its high conduction closes the low resistance path, which causes network Z1 to discharge. The electron flow, which takes place through coil L3 during discharge of network Z1, would ordinarily cause V3 to conduct. However, the short duration of the pulse (approximately 0.25 microsecond), combined with the 6.5-henry inductance of L1, presents a large impedance which prevents the pulse current from being partly reversed through the power supply. Pulse-forming network Z1, which is considered a delay line, is provided with a bleeder series resistance to protect maintenance men (the large value of this resistance does not affect operation of the network). The inductances shown in the network are used to steepen the sides, flatten the top of the output pulse, and to increase the discharge time to approximately 0.25 microsecond. The complete operating cycle of the pulse-forming network lasts 1,000 microseconds; discharge time, approximately 0.25 microsecond; charge time, about 400 microseconds; time charged, about 599.75 microseconds. The discharge of network Z1 is a 6,500- to 8,500-volt, 0.25-microsecond pulse, which is applied through jack J3 and the slip-ring assembly to the primary of the pulse transformer.
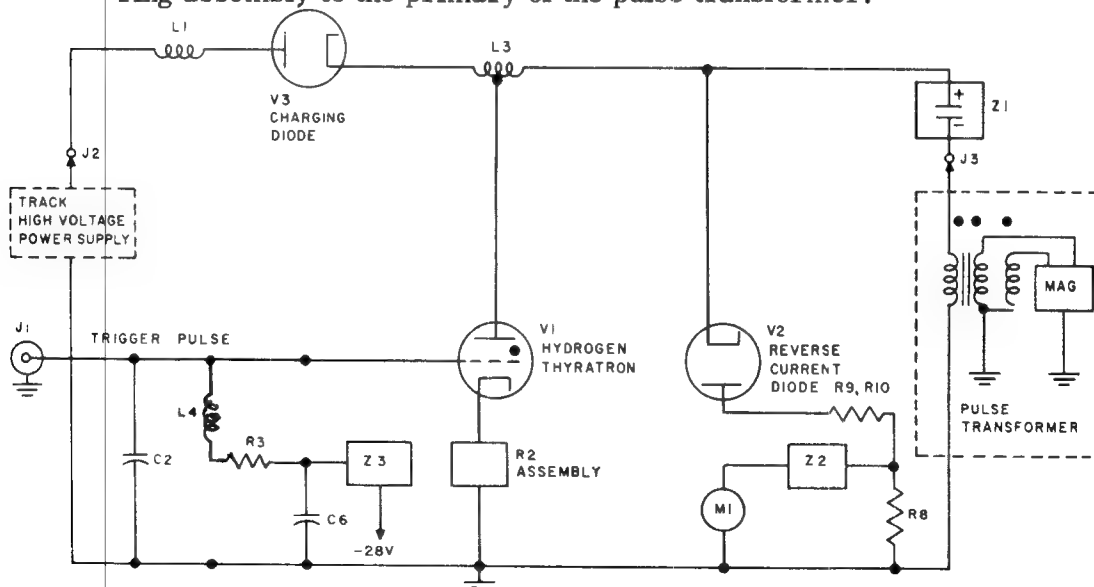

Figure 6. Track modulator, simplified schematic.

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(3)   Reverse-current discharge (fig 6). Although the 50-ohm characteristic impedance of network Z1 matches the 50-ohm impedance of the coaxial cable and the primary of the pulse transformer, sporadic arcing of the magnetron during operation may cause an occasional mismatch. In this case, the increased reverse current which results not only neutralizes the positive charge on network Z1, but also places a negative charge on network Z1. Reverse-current diode V2 then conducts and prevents the negative charging of network Zi. Electron flow is from the upper plates of the Z1 network capacitors, through V2, resistors R8, R9, and R10, filter network Z2, meter M1, through ground, the primary of the pulse transformer, and to the lower plates of the capacitors. The shunting action of V2 allows network Z1 to begin charging from zero potential, so that the output trigger is always at a constant level. The amount of mismatch caused by improper operation of the magnetron causes a proportional amount of current to flow through resistors R8, R9, and R10. Part of this current is taken from the junction of resistors R8 and R9 and is applied to meter M1, which provides a visual indication of the mismatch. Since the duration of the reverse-current discharge is very short, capacitor C1 is connected across resistors R8 and R9 to prolong the time of current flow through the meter so that it can provide an indication. Capacitor C1 charges during the time of current flow and discharges through the resistors when the reverse-current flow stops. Impedance network Z2 provides filtering for the current applied to meter M1. It might appear as though the reverse-current discharge would flow through diode V3 and inductance L1, to the power supply. However, the short time of discharge and the inductance of L1 present a large impedance, compared to the resistance offered by tube V2, resistances R8, R9, and R10. Consequently, the reverse-current discharge through inductor L1 is negligible.

(4)   Thyratron input circuit. Capacitor C2 prevents accidental firing of thyratron tube V1 (when this unit is used in the missile-tracking radar) by the discharge of the missile track modulators (TM 9-5000-21). Each time one of the two modulators in the missile-tracking radar discharges a high-voltage pulse to the magnetron, the pulse appears in the other modulator across the series cathode-to-grid and grid-to-plate capacitance of thyratron V1 and across the 0.5-ohm resistor in the R2 assembly. If capacitor C2 were not connected between the grid of V1 and ground, most of this high-voltage surge would appear across the small cathode-to-grid capacitance. This voltage would be enough to ionize thyratron V1. With capacitor C2 (which has a higher capacitance value than the cathode-to-grid capacitance) connected between the grid of V1 and ground, most of the high-voltage pulse appears across the

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grid-to-plate capacitance. This feature prevents thyratron V1 in one of the missile-tracking radar modulators from being accidently fired by the pulses of the other missile track modulators. Capacitors C2 and C6 and inductance L4 are used in the target-tracking radar to help preserve the waveshape of the input pulse. The distributed capacitance of L1 and inductance of L1 cause spurious oscillations which are fed back to the thyratron grid through the power supply. These oscillations act to prolong the conductivity of the thyratron after the end of the trigger pulse. To counteract this effect and to prevent accidental firing of the thyratron, a fixed bias of -28 volts is placed on the thyratron grid, through impedance network Z3, resistor R3, and inductance L1. Inductance L4 presents a large impedance to ground to the trigger generator pulses. In this manner, the signal is fully applied to the thyratron grid. Capacitor C6 provides an r-f ground (decoupling) for these pulses.

(5) Miscellaneous component usage. Modified sets use the output of test point one (TP1) to provide, in effect, a sync signal to the range unit. This arrangement allows the range unit to know when the magnetron has fired and clocks down to determine range. Blower motor B1, with its associated circuitry, prevents hot spots that might appear within the unit. The blower has a single-phase, 120-volt, 400-cycle motor which turns at approximately 2,500 revolutions per minute (rpm). Interlock switch S1 is closed, and interlock switch S3 is open when the cover to the modulator is removed. Switch S3 breaks the power interlock circuit and removes voltage from the primary of the track modulator high-voltage power supply. Switch S1 shorts the output of the power supply entering the track modulator to ground to remove the charge across the filter capacitor in the power supply. This action of S1 prevents damage to the power supply and affords protection for maintenance men.

11. SLIPRING PULSE TRANSFER (fig 3-7)

The output from the track modulator is transferred to the pulse transformer through two of the rings in slipring assembly GS-15536. This procedure is used because the track modulator is mounted on the stationary portion of the target-tracking radar antenna trailer, while the pulse transformer and the magnetron are mounted on the rotating antenna. Sliprings A and B transfer the pulse. They are separated from the rest of the sliprings for voltage insulation purposes. The sliprings do not effect the power or the shape of the high-voltage pulses. Coaxial cable conducts the pulse from the track modulator to the sliprings, and from the sliprings to the pulse transformer. Six pairs of contacts are used on slipring A, the high-voltage ring. Slipring B is for ground return.

12. MAGNETRON OSCILLATOR (TRACK)

a. General (fig 3-8). The X-band 5,780 magnetron target-tracking r-f unit GS-15598 oscillates at a frequency variable between 8,500 and 9,600 megacycles. The peak power output is from 200 to 400 kilowatts over the operating frequency range. The magnetron oscillates for 0.18-microsecond periods because of the trigger pulses from track modulator GS-15461. These pulses are coupled through the windings of the pulse transformer and are applied to the magnetron. The high-energy r-f produced by the magnetron is conducted by waveguide to the r-f and antenna system. The magnetron current is monitored by MAG CURRENT meter M1 in equipment panel B and also by MAGNETRON HV SUPPLY meter M1 in target track control panel GS-15584. After the magnetron filaments have warmed up with full heater current and 5-minute delay timer GS-16041 has reduced the heater current to a standby value, the magnetron is ready to oscillate. Magnetron current energizes MAG CURRENT relay K4 in equipment panel B, which places an additional series resistance in the filament circuit to further reduce its voltage and prevent overheating because of magnetron oscillations. The action of the same relay (K4) applies power to the shutter circuit, which removes the shutters from the waveguides. The magnetron is tuned over its frequency range by a magnetron tuning drive. A metering circuit is provided to indicate the operating frequency of the magnetron (meter located on the target control panel).

b. Magnetron heater. To prevent the magnetron from drawing current before the filaments are fully heated, a 5-minute delay circuit is provided in the power supply to heat the filaments quickly and to prevent the track modulator high voltage from being immediately applied to the magnetron (fig 7). Transformer T3 is the magnetron filament transformer. MISSILE-TARGET switch S4 in equipment panel B is for inserting different resistance values in the heater circuit for correct operation of the magnetron when the unit is used in the missile-tracking radar. MISSILE-TARGET switch S4 is in the TARGET position for the following discussion. When the target-tracking radar power is turned on, 120-volt, 400-cycle power is applied through HEATER INTLK switch S6 and fuse F1 to the 5-minute delay timer. Action in the timer causes 5-MIN DELAY relay K1 in equipment panel B to energize (ST 44-161-2m). Full 120-volt, 400-cycle power is then applied to the primary of transformer T3. The energizing path is as follows: from a-c neutral at E41-7, through contacts 5 and 6 of deenergized relay K4, through contacts 2 and 3 of energized relay K1 through transformer T6 to terminal 1 of transformer T3, and then through switch S6 to the 120-volt source. For protection of maintenance men, HEATER INTLK switch S6 breaks power to the 5-minute delay timer and to transformer T3, when the door to the hot box is opened. Hot box is the common term used when referring to the part of the transmitter assembly containing pulse transformer

T4. Five minutes after the target-tracking radar power is turned on, the 5-minute delay timer deenergizes relay K1. This inserts resistor R1 in series with one side of the primary of transformer T3, effectively reducing the filament current in the magnetron. The magnetron filament is now in standby, having enough voltage applied to remain heated. The 5-minute delay timer also prevents high voltage from being applied to the magnetron from the modulator during the preheat period. At the end of the 5-minute delay period, the target-tracking radar operator can apply the high voltage to the modulator unit which then pulses the magnetron. The magnetron draws current and oscillates. This magnetron current energizes MAG CURRENT relay K4 in equipment panel B, which places resistor R2 in series with resistor R1 and one side of the primary of transformer T3. The increase resistance further reduces the magnetron filament current, thus preventing the filament from overheating when the magnetron is operating. For a comprehensive coverage of the high-voltage interlock and control circuitry, refer to TM 9-5000-17. Transformer T6 is inserted in series connection with one side of the filament transformer to raise the magnetron filament voltage during the preheat period.

c. R-F generation (fig 7).

(1) The 7,500-volt negative pulse from the track modulator is shaped by capacitor C18 and choke L2 and stepped-up by pulse transformer T4. The secondaries of transformer T4 then apply a negative pulse (with respect to ground) of 30 to 35 kilovolts, to the filament-cathode circuit of the magnetron. Since the plate of the magnetron is connected to ground, the large negative pulse on the cathode of the magnetron causes it to conduct and oscillate. The trigger pulse across the primary of pulse transformer T4 is modified by bifilar inductances L1A and L1B to improve the magnetron power output and current pulse form. If the full trigger voltage were suddenly applied across the magnetron, it would arc and operate sporadically. To avoid this, it is necessary to gradually apply the trigger pulse voltage until the magnetron is able to oscillate enough to support the full voltage across it. These inductances place a step on the leading edge of the trigger pulse at the firing point of the magnetron, giving the magnetron a chance to start oscillating before full voltage is applied. Inductances L1A and L1B combined with distributed capacitance tend to oscillate so that the first and the last quarter-cycles of operation are out of phase, while the intervening half-cycle is in phase with the trigger pulse. This action shapes the trigger pulse so that optimum operation is obtained from the magnetron. The leading edge of the pulse entering the magnetron has a light step in it, and the trailing edge of the pulse drops sharply. Resistors R28 through R30 are used to reduce the Q of inductances L1A and L1B to prevent an undesirable effect. Capacitors C9, C10, C11, and C12

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assure that no high-voltage difference appears between filament leads during trigger pulses. Capacitors C13 and C14 provide an r-f short to ground, thereby preventing the high-voltage pulses from entering network Z1 and the metering circuits. Resistor R44 gives additional protection for network Z1 and the metering circuits.

Figure 7. Magnetron and control circuits, simplified schematic.

(2) When the magnetron is fired by the trigger pulse, electron flow takes place as follows: From terminals 6 and 4 of transformer T4, through L1A and L1B to the cathode, to the plate of the magnetron, to ground; through series-connected meters M1 (MAGNETRON-HV SUPPLY) in the target track control panel GS-15584 and MAG CURRENT meter M1 in equipment panel B, from ground, through resistor R4 in equipment

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panel B, through MAG CURRENT relay K4; through impedance net-
work Z1; through resistor R44; to terminals 3 and 5 of transformer T4.
MAG CURRENT relay K4 acts to reduce magnetron heater current
after the magnetron draws sufficient current. (See discussion of
magnetron heater.) Energized MAG CURRENT relay K4 also applies
power to the shutter circuit, which in turn removes the shutters from
the waveguides. The removal of the shutters allows returned signals
to be applied to the receiving system. Further discussion of these
shutters will be found by referring to chapter 5, paragraph 24a(2) of
this text.

d. Arc suppressor. To prevent the magnetron from arcing excessively, an
arc suppressor circuit has been provided. The circuit consists of arc sup-
pressor relay K6 and capacitors C19 and C20 (fig 8). One side of the relay is
connected to +150 volts through pin 16 on terminal strip E42 and the other side,
to the arc suppressor electrode. Capacitor C20 keeps the sharp pulses caused
by extremely short arcs out of the 150-volt supply. Capacitor C19, shunting
relay K6, prevents relay K6 from being energized by arcs of short duration.
For arcs sufficiently long for capacitor C19 to be charged, relay K6 will
energize. This interrupts the high voltage (by energizing HV ON relay K5)
until the arc is extinguished and capacitor C19 has discharged through the arc
suppressor relay coil. Capacitor C19 (1μfd) and the winding resistance of
relay K6 provides a time constant which is long enough to permit the magnetron
to clear itself before relay K6 deenergizes.

e. Monitoring and protection.

(1) Meters are used to monitor the magnetron current for two reasons:
First, optimum operation of the magnetron is indicated by its amount
of cathode-to-plate current; and second, any malfunction of the mag-
netron during operation is indicated by abnormal magnetron current
readings. Two meters are connected to the cathode circuit of the
magnetron to indicate magnetron current (figs 7 and 8). MAG CURRENT
meter M1 in equipment panel B is used to monitor the current locally
at the magnetron, and MAGNETRON-HV SUPPLY meter M1 in the
target track control panel GS-15584 monitors magnetron current at the
location of the radar operator. The meter in the control panel is lo-
cated directly above variable transformer T3. This transformer
controls the output of the high-voltage supply to the modulator and
consequently the amplitude of the pulses applied to the magnetron. By
use of variable transformer T3, the current for optimum magnetron
operation may be obtained and observed on MAGNETRON HV-SUPPLY
meter M1. Impedance network Z1 in the magnetron cathode circuit
is used as a filtering circuit to keep r-f variations out of the metering

circuits. The filtering action of this network maintains a constant reading on the meters from pulse to pulse.
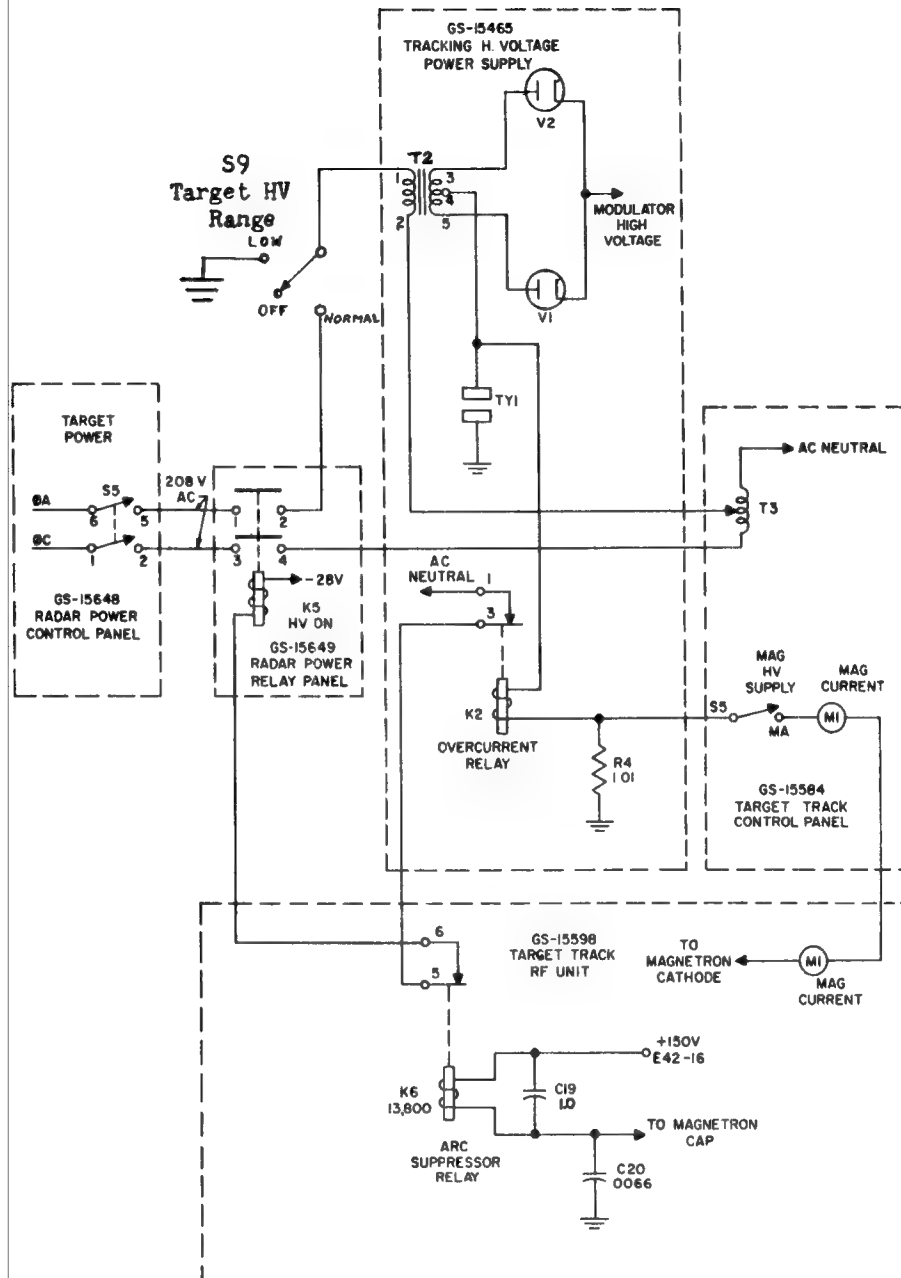

Figure 8. Arc suppressor high-voltage cutoff, simplified schematic.

(2) Should the magnetron current somehow be prevented from flowing through the tube, in such a manner that a large voltage is built up on the

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secondaries of the pulse transformer, this voltage discharged through spark-gap E47 so that the transformer is protected from overvoltage.

(3) When the magnetron is operating, producing a large amount of heat, a cooling system consisting of a blower motor and an air duct around the magnetron is used. Blower motor B1 is a three-phase induction motor using all three phases of the 400-cycle power. When the main power switch for the target-tracking radar is closed, this motor is activated. In case the motor fails to operate, blower switch S5 breaks the target interlock circuit removing power from the track modulator high-voltage supply, and thus removing high voltage from the magnetron. Switch S5 is kept closed only by the air stream from the blower. The MAIN DOOR SWITCH-RESTORABLE switch S3 also breaks the target interlock circuit when the door to the r-f unit is opened. This switch may be manually reset with the power on for maintenance work on the r-f unit.

f. Magnetron tuning drive.

(1) The frequency of the 5,780 magnetron can be remotely controlled from the target track control panel GS-15584 in the target track console. It can also be locally controlled from equipment panel A in the target track r-f unit GS-15598. The magnetron frequency is adjusted by varying the size of its resonant cavities, and thus, its resonant frequency. This adjustment is accomplished by sliding tuning slugs in or out of these cavities. The slugs are alined so they slide inside the cavities without shorting to the sides (fig 9).

(2) For remote tuning, FREQ DECREASE toggle switch S1 and FREQ INCREASE switch S2 in equipment panel A are in their normal position, thus connecting lines Mag F1 and Mag F2 directly to motor B2 in the BR-710993 magnetron tuning drive. Lines Mag F1 and Mag F2 are connected to FREQUENCY INCREASE-DECREASE switch S6 in the target track control panel GS-15584. When this switch is pushed to the left (DECREASE position), a-c neutral is connected through capacitors C3, C4, and C5 in the equipment panel to terminal 1 of magnetron tuning drive motor B2, and 120-volt phase-A power is applied to terminal 5 of motor B2. Terminal 3 of motor B2 is permanently connected to 400-cycle a-c neutral. In the DECREASE position of switch S6, motor B2 turns so that the magnetron tuning slugs tend to be pulled out of the cavities. When switch S6 is pushed to the right (INCREASE position), 120-volt phase-B power is applied to terminals 1 and 5 of magnetron tuning drive motor B2. This causes motor B2 to reverse direction and to insert the tuning slugs into the cavities. When motor B2 rotates, it turns a worm gear connected to the calibrated tuning head of the

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magnetron. The tuning head turns a screw which slides a block holding the tuning slugs in or out of the magnetron cavities, depending on the direction of rotation of motor B2. This action causes the cavities to be smaller or larger, thereby raising or lowering the frequency at which the cavities resonate.
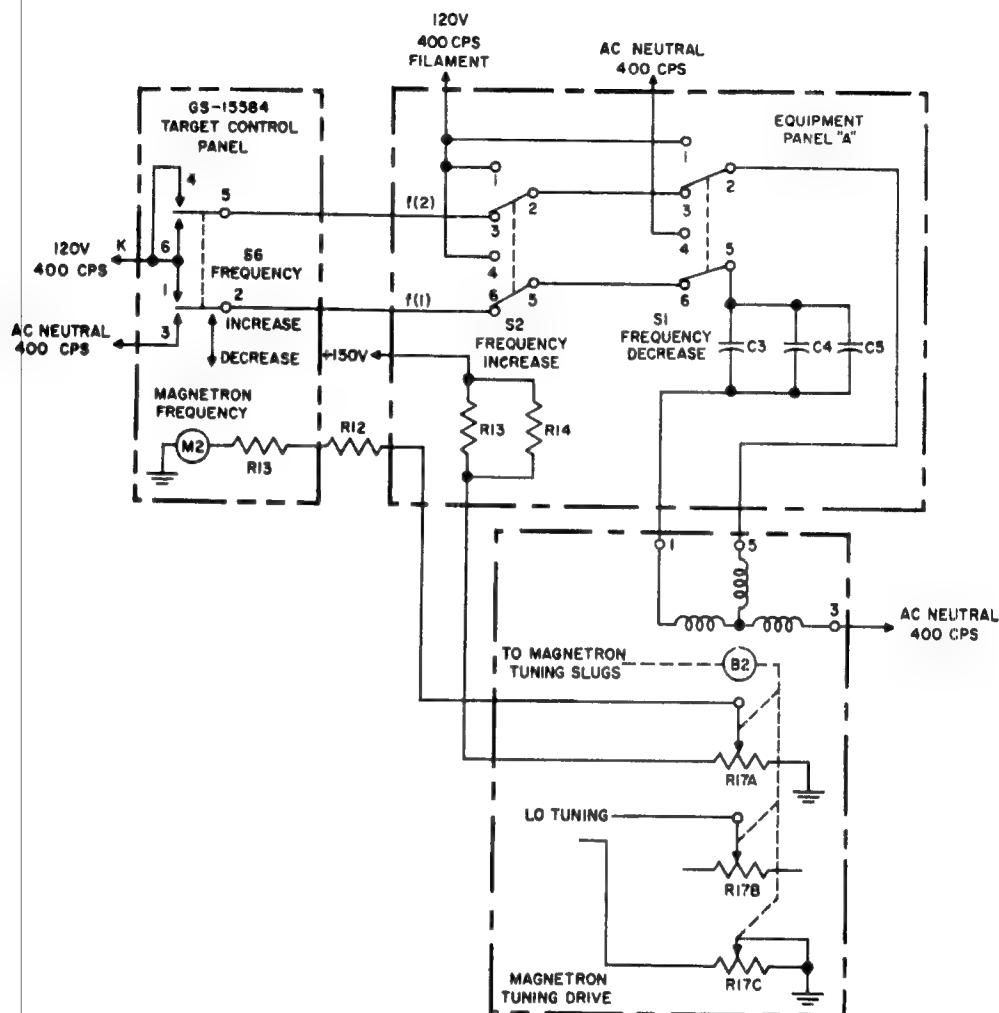

Figure 9. Magnetron frequency control circuit, simplified schematic.

(3) For local control of magnetron frequency, either switch S1 (FREQ DECREASE) or switch S2 (FREQ INCREASE) in equipment panel A is used. These switches are spring-loaded and have to be manually held in the operating position. When FREQ INCREASE switch S2 is pushed,

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120-volt phase-B power is applied to terminal 1 of motor B2, and 120-volt phase-A power is applied to terminal 5 of motor B2. This causes motor B2 to turn in such a direction as to pull the slugs out of the cavities.

(4) Frequency indication of the magnetron is provided by FREQUENCY meter M2 in the target track control panel GS-15584. The meter is calibrated to read the approximate operating frequency of the magnetron. When tuning motor B2 turns, it moves the taps of potentiometers R17A, R17B, and R17C in the BR-710993 magnetron tuning drive. Potentiometer R17A and resistors R13 and R14 (equipment panel A), make up a voltage divider from +150 volts to ground. The voltage tapped off potentiometer R17A is proportional to the magnetron frequency, and thus this frequency is indicated on FREQUENCY meter M1 in the target control panel. Potentiometers R17B and R17C are used to change the local oscillator frequency in the receiving system when the magnetron frequency is changed. This point is discussed in chapter 5, paragraph 24a(5).

g. Magnetron aging. In order to age magnetron tubes which are gassy and will not stably operate at the lowest setting of the MAGNETRON-HV SUPPLY control, TARGET HV RANGE switch S9 has been added in radar power cabinet assembly GS-15515. Refer to figures 8 and 9. This switch has two sections; one section is in the high-voltage supply circuit, the other, in the low-voltage interlock circuit. For ordinary operating conditions, switch S9 is set in the NORMAL position (fig 8). This connects 208 volts a-c across the primary of transformer T2 in the tracking high-voltage power supply GS-16565 and a variable portion of the MAGNETRON-HV SUPPLY control transformer T3 on the target track control panel GS-15584. In the OFF position, switch S9 opens the a-c supply to transformer T2 and also the HIGH VOLTAGE OVERCURRENT relay K2. This in turn causes the HV ON relay K5 to open and remove power from the magnetron. The LOW position of switch S9 connects a-c neutral to terminal 1 of transformer T2 and applies 120 volts across a variable portion of transformer T3 and the primary of transformer T2. The switch cannot be turned from NORMAL to LOW or LOW to NORMAL without momentarily releasing pressure on the switch in the OFF position. This feature is included as a safety measure to assure opening of the interlock circuit. The interlock circuit (fig 10) remains open because the HV ON relay K9, which is normally locked in, is deenergized when switch S9 is turned to OFF. After TARGET HV RANGE switch S9 is set to LOW or NORMAL and the target track high-voltage (START-MAX) control is turned back to START, the HV ON relay K9 again locks in and closes the interlock circuit. A-C neutral is connected through the interlock circuit to pin 1 of OVERCURRENT relay K2 (fig 8) and power is

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reapplied to the magnetron through the contacts of HV ON relay K5. This
sequence of operation protects the magnetron from high voltage by preventing
the application of magnetron power while the target track high-voltage (START-
MAX) control S2 is turned to a high setting. The principal reason for magnetron
aging is to clean up the small amounts of gas that evolve during periods of
nonoperation. The period required for aging will vary from tube to tube and
from transmitter to transmitter, and usually is a function of the time the
magnetron has been inoperative. For initial aging of the magnetron, the pulsed
voltage should be applied gradually, allowing the tube to arc a little at each
higher level until the arcing ceases. It is further recommended that this aging
be carried to levels slightly (not more than 10 percent) higher than the normal
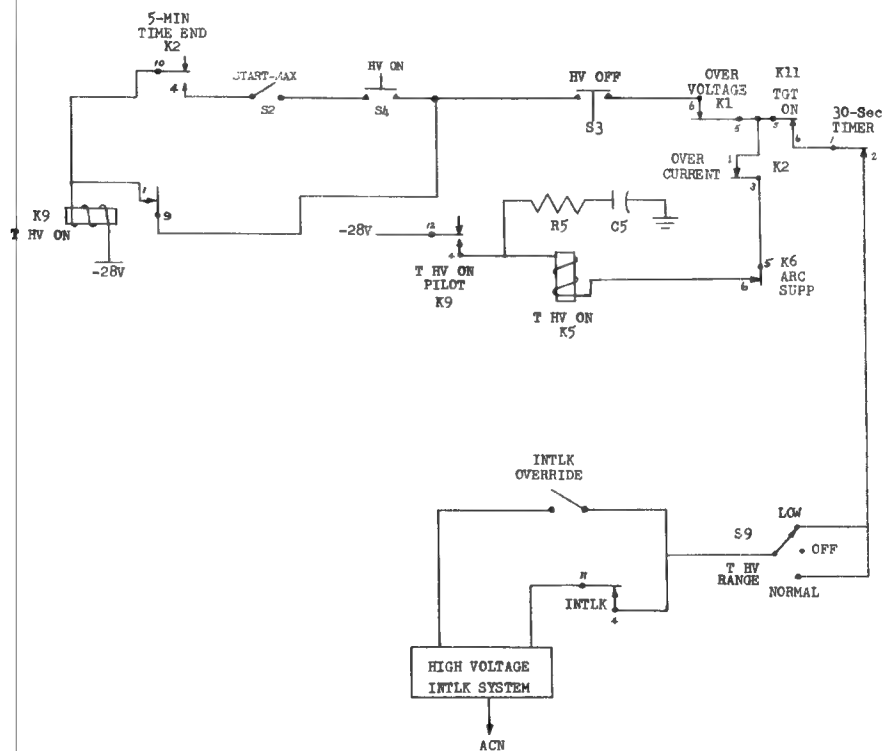operating point of the transmitter to obtain optimum stability.

Figure 10. Low-voltage interlock circuit, simplified schematic.

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CHAPTER 4

R-F AND ANTENNA SYSTEM

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13. GENERAL

a. The r-f and antenna system radiates the r-f energy pulses into space from the transmitting system and receives the returned signals of these pulses reflected by a target. The direction finding and ranging operation is carried out by the r-f and antenna system using the monopulse principle described in paragraph 14, Introduction to Monopulse Radar. The r-f pulses are emitted in an extremely narrow beam by a lens-type metal antenna. The reflected pulses received at the antenna are applied to the monopulse waveguide plumbing which supplies three signals containing information necessary for ranging and automatic tracking of the moving target. The three signals applied to the receiving system are: (1) total strength of the received signal (or amplitude sum), (2) azimuth difference, and (3) elevation difference.

b. Another function of the r-f and antenna system is the transmit-receive switching used in coordinating the actions of the transmitting and receiving systems. The r-f and antenna system also provides sampling of the transmitted pulse that is used for automatic frequency control in the receiving system for power monitoring. The r-f and antenna system comprises the track waveguide assembly GS-15600, the track lens assembly GS-15645, four 6163 ATR boxes (tubes), three 6164 TR boxes (tubes), and two directional couplers. All of these components are located in the target track r-f unit GS-15598.

c. The r-f pulses from the 5,780 magnetron in the transmitting system are coupled by the waveguide to the four ATR tubes. These tubes permit easy conduction of the pulses to the waveguide plumbing of the antenna assembly. Refer to figures 3 through 8 and to simplified schematic, figure 11. The three TR tubes located between the receiving system and the antenna system keep the high-power transmitting pulses from entering the receiving system. The pulses travel through a complex network of hybrid tee junctions and waveguides to the four antenna radiating horns. The radiating horns, with the aid of the antenna, produce a pencil beam of radiated energy having a beam width of 1.2° between the half-power points. Therefore, most of the energy during transmission is concentrated on the target. During reception all the energy impinging upon the lens area is focused into the relatively small area presented by the opening of the radiating horns, since these horns are located at the focal point of the lens.
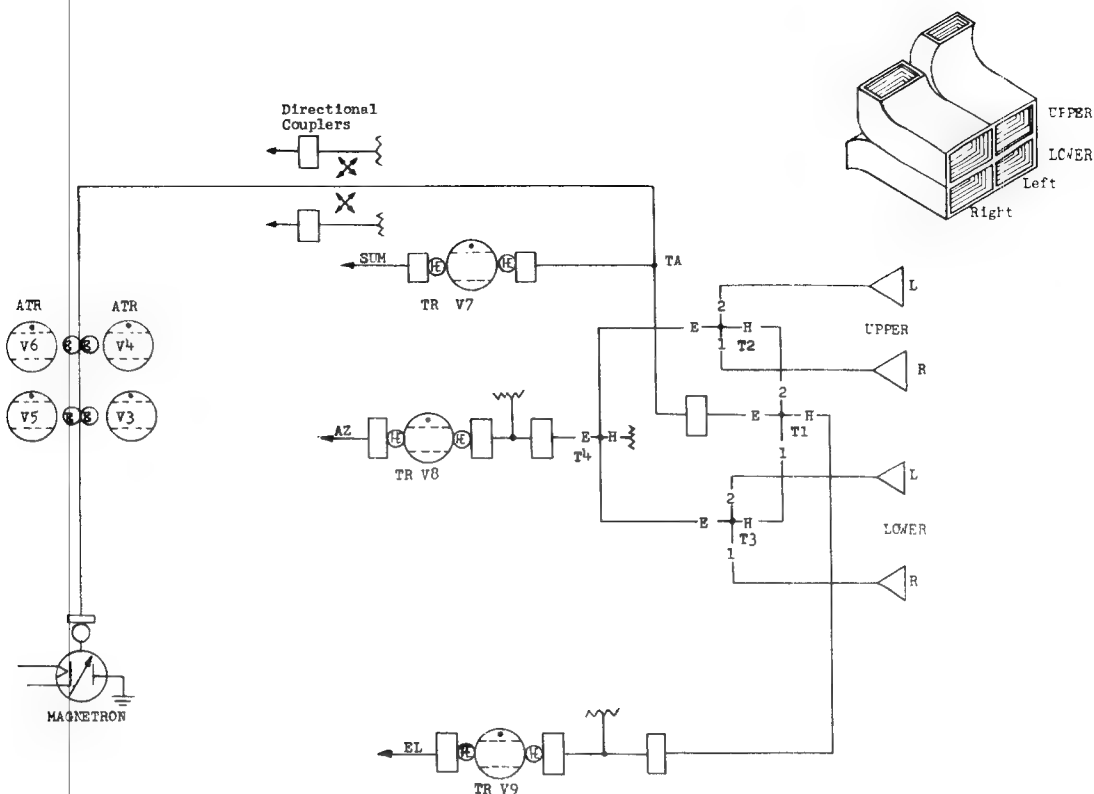
Figure 11. Track waveguide assembly, simplified schematic.

14. INTRODUCTION TO MONOPULSE RADAR

a. All radars are essentially radio direction finders. The fire control radar differs from the acquisition or search radar only in the greater accuracy with which it determines the direction (azimuth and elevation) and range. With monopulse radar (such as Nike tracking radars) signals representing azimuth and elevation angular errors as well as signals for range determination are obtained. The main difference between monopulse radar and conventional fire control radar lies in the operation of the direction finder.

b. Radars that use the lobing system point the r-f beam first to one side of the target and then to the opposite side comparing the strength of the returned echoes between the different antenna positions. The difference in strength or amplitude between the returned signals supplies an error voltage to the servo system which turns the antenna so as to correct the error. Therefore, when opposite pulses of amplitude are equal, an on-target position exists.

c. In the monopulse system the direction information is obtained from a single pulse instead of several pulses, as in the lobing system. The monopulse system uses four antennas and compares the amplitude of the portions of a single returned pulse as it is distributed into the four antennas. In the Nike system, four flared pieces of waveguide forming radiating-horn antennas are arranged in a square at the focal point of the metallic lens antenna. During transmission of a pulse, each of the four horns radiates the same amount of energy, in the same phase. During reception, if the antenna is pointing off-target signals of different amplitude will enter the four radiating horns and will be mixed in the monopulse system of waveguides. This mixing yields error signals in azimuth and elevation, provided the antenna is pointing off target in azimuth and elevation. The error signals are applied to a servo system which moves the antenna to the on-target position. When an on-target condition exists, the portions of returned signals are of equal amplitude and no error results from their mixing. The mixing of signals will always yield a signal proportional to the sum of their amplitudes. This signal is used for range determination.

15. MONOPULSE DESIGN PROBLEMS

a. General. The accuracy of pointing-error information in a monopulse radar is dependent upon certain factors not readily apparent; in particular, three factors: relative attenuation in the waveguide assembly before amplitude comparison; relative phase shift in the waveguide assembly before amplitude comparison; and relative phase shift in the three receiver channels after amplitude comparison. In order to understand the effects which these three factors have upon the accuracy of the system, a knowledge of certain properties of the angle error detectors is necessary. These detectors, one for azimuth and one for elevation, receive both the sum-channel signal and the appropriate difference-channel signal (azimuth signal or elevation signal) and operate to produce a d-c output voltage with a magnitude proportional to the pointing error. The angle error detector makes a phase comparison between its input signals. The polarity of its output voltage after rectification in the error pulse rectifier, which is dependent upon this comparison, indicates the direction of the pointing error. As will be shown in the information on the angle error detectors, the output voltage of the detector stage is

$$E_{out} = KE \cos \emptyset \qquad\qquad (1)$$

where K is a constant, E is the input from the difference channel, and $\emptyset$ is the phase angle between the inputs applied from the sum channel and the difference channel. Because the cosine of 90° is zero, a phase relationship of 90° between the two inputs will cause the output voltage to be zero. However, the output of the angle error detector must be zero only when an on-target condition exists.

Hence, it is necessary that the phase relationship of the two input signals to the angle error detector pass through 90° only at the crossover point, that is, when the target is at the center of the beam. In an ideal system, the two inputs are in phase when the target is at one side of the beam and are of opposite phase when the target is at the other side of the beam. An abrupt reversal in the relative phase of the two signals occurs as the target passes through the center of the beam, and at that instant the magnitude of the difference channel signal is momentarily zero. In a practical, nonideal system the phase reversal is not instantaneous and the magnitude of the difference-channel signal never goes to zero. The phase relationship reverses in the vicinity of the crossover point. To avoid a pointing error, the changing phase relationship must pass through 90° exactly at the crossover point. This fact must be kept in mind during the discussion which follows.

b. <u>Relative r-f attenuation.</u> When on target with an ideal system, the inputs to hybrid junction T3 of figure 11 are exactly equal and 180° out of phase. Cancellation of these inputs takes place in the H arm and no signal is present in the elevation channel of the receiver. In a practical system, unavoidable physical difference will exist between the paths followed by the two inputs to junction T3. These differences may cause the signal in one path to experience a greater over-all attenuation than is experienced by the signal in the other path. Because of this difference in attenuation (gain symmetry), complete cancellation will no longer take place in the H arm and a difference signal will be present. This signal, after passing through the elevation channel of the receiver, will ultimately cause the antenna to be positioned off target by the amount necessary to produce a zero r-f elevation error signal. The resultant error is not troublesome because it is a systematic error of the equipment and is compensated for in the collimation procedure.

c. <u>Relative r-f phase shift.</u> In a practical system, physical differences will also cause a difference in the over-all phase shift experienced by the two inputs to hybrid junction T3. (Although junction T3 is being used in this explanation, the principles involved are applicable to the remaining components.) Because of this phase asymmetry, the two inputs to T3 will not be exactly opposite in phase and complete cancellation will not take place in the H arm. This incomplete cancellation differs from that caused by gain asymmetry in that cancellation cannot be obtained by changing the position of the antenna to produce a change in the relative magnitude of the two inputs. This may be seen from an inspection of the vector diagram of figure 12. It is apparent that a change in the magnitude of either input will not serve to eliminate the signal present in the H arm. Hence, phase asymmetry prevents the r-f error signal from becoming zero. Instead, a minimum signal (null) is obtained when the two inputs to the junction are of equal magnitude. One effect of obtaining merely a null when on target, rather than zero, is to reduce the sensitivity with which it is possible to detect the on-target

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condition. For example, consider the case of a target flying through the center of a stationary beam. The magnitude of the r-f error signal decreases as the target approaches the center of the beam, passes through minimum (null) as the target passes through the beam axis, and increases as the target moves away from the center of the beam. If only slight phase asymmetry is present the null will be deep, and the crossover point will be easily detected at the instant when the phase of the H arm signal reverses sharply. If considerable phase asymmetry is present, the null will be shallow and the crossover point will become obscured. It is also apparent from figure 12 that the phase relationship between the sum and difference signals at the output arms of T3 is 90° when, and only when, the two inputs are of equal magnitude. In the absence of later phase shifts, the phase relationship of the two signals at the input to the angle error detector will be 90° at the instant when the target is on the beam axis, and no pointing error will be introduced.
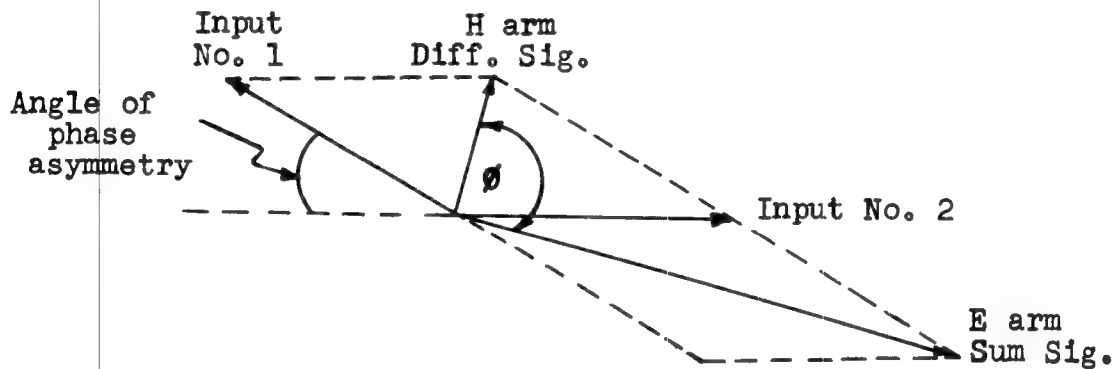
Figure 12. R-F phase asymmetry, junction T3.

d. Relative i-f phase shift. After amplitude comparison in the hybrid junction of the waveguide assembly, the sum signal and the two difference signals pass through separate channels of conversion and i-f amplification. The components contained in these separate channels introduce an over-all phase shift of approximately 15,000°. Following their passage through these channels, the difference i-f signals are compared in phase with the sum i-f signal. Any inequality in the over-all phase shift introduced in these channels will cause the phase relationship at the point of comparison (angle error detector) to differ from that which existed at the output of the hybrid junctions. In a system in which no phase asymmetry is present prior to the hybrid junction, the difference output of the hybrid junction is zero at crossover. In that case, obviously no amount of phase shift throughout the remainder of the system could change the position of the zero null, and the output of the angle error detectors would be zero at crossover. In a practical system, a difference signal has a 90° phase relationship to the sum signal. Unequal phase shift in the remainder of the

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system would cause these two signals to have a phase relationship other than 90° at the angle error detector. The angle error detector would then have an output voltage when the system was actually on target.

16. TRANSMIT-RECEIVE SWITCHING

a.   The monopulse radar system, which uses the waveguide assembly as a common channel for transmission of the high-power r-f pulses and for reception of the weak reflected pulses, requires a switching arrangement whereby the receiving system presents an infinite impedance to the transmitted pulse. The transmitting system presents an infinite impedance to the received pulse. In this manner, the crystals of the receiving system (balanced converter) will not be damaged by the high-power pulses, while the entire strength of the weak received signal pulses will be fully applied to the receiving system.

b.   The switching operation is accomplished by means of TR (transmit-receive) tubes and ATR (antitransmit-receive) tubes. The construction of either type is fundamentally the same. They are gas-filled tubes containing two conical electrodes which provide a spark-gap when high-power pulses cause the gas to ionize. The TR tubes, in addition, have a connection for the keep-alive voltage discussed in paragraph 21 of this chapter. Figure 13 is a simplified drawing of TR and ATR tubes showing their connection with the main waveguide and wavelengths (λ).

c.   During transmission the high-power pulse entering the ATR box ionizes the tube whose electrodes are located at one-quarter wavelength ($\lambda/4$) from the main waveguide. Firing the spark-gap is very nearly equivalent to closing the aperture where the ATR tube joins the main waveguide. This tube maintains the entire amount of r-f energy flowing along the main waveguide to junction A (figure 11). At this point, part of the high-power pulses will continue to the antenna and part will enter the receiving channel. However, the TR tube, located at $\lambda/4$ from the junction, will ionize and cause the same effect as the ATR tube, thus reflecting the high-power pulse back to the main waveguide.

d.   During reception the weak incoming pulse will not be sufficient to ionize the TR or ATR tubes. At junction A part of the pulse will be coupled to the the transmitting channel and will enter the ATR box. This time, however, the tube will act as an open circuit and the reflected energy will appear as an infinite impedance at the center of the main waveguide ($\lambda/2$ from the ATR tube). The same effect will be present at the A junction located an even number of quarter wavelengths away from the ATR tube. Consequently the entire strength of the received pulse will go through the TR tube to the receiving system.

e.   Figure 11 shows the four ATR tubes V3, V4, V5, and V6. Tubes V3 and V4 were first used, but the impedance of the waveguide was changed at the

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junction of the waveguide and magnetron. Because this change of impedance caused some arcing at the magnetron window, tubes V5 and V6 were added to counteract this characteristic. TR tubes V7, V8, and V9 are located at the three inputs to the receiving system.
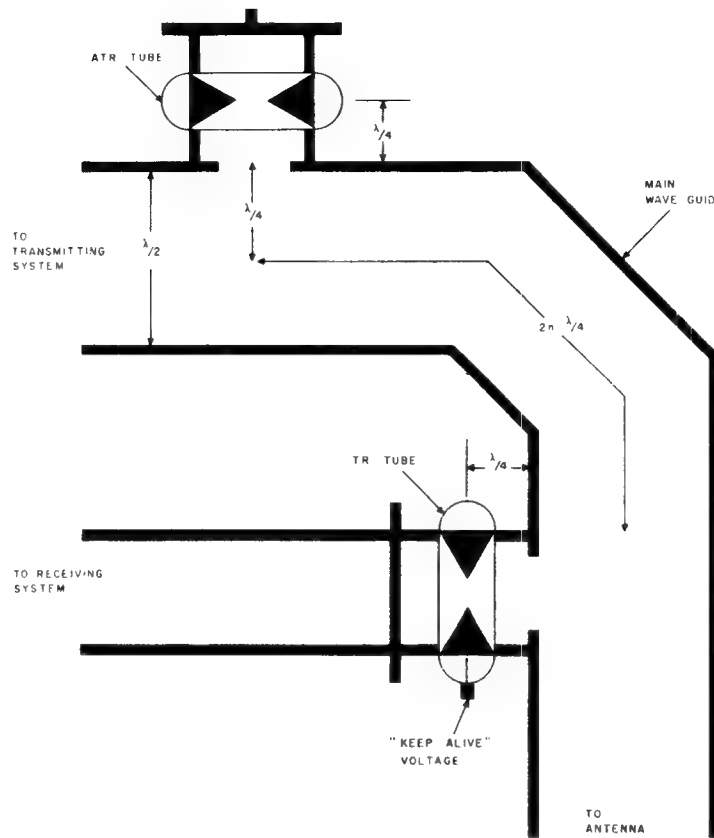
Figure 13. Transmit-receive switching.

17. HYBRID TEE JUNCTION (fig 14)

a. The monopulse system of signal distribution during transmission and signal mixing during reception is accomplished with the combination of four hybrid tee junctions. A hybrid tee junction can be described as a dual electric-magnetic-plane type of tee junction. The plane described by arms 1-H-2 of the tee is the magnetic or H plane of the tee. The plane described by arms 1-E-2 is the electric or E plane of the tee. The cross-sectional view of an electric-plane junction in figure 15 shows lines of electric intensity drawn in successive positions of the same wavefront entering arm E. This figure indicates the result of the waves reaching the junction point and the phase relationship between the output of arms 1 and 2. Although at the junction point there

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will be some reflection (not indicated), the input wave power tends to be divided equally between arms 1 and 2 if the planes AA' are 180° out of phase with the waves at BB' . In figure 16 wavefronts enter the electric-plane junction from arms 1 and 2. If these two entering waves are equal in phase and amplitude at planes AA' and BB' , when they reach the E arm they will be 180° out of phase and will cancel each other. If the two electric waves entering arms 1 and 2 are different in amplitude but equal in phase, their amplitude difference will appear in the E arm, with the resultant arrows pointing in the direction of the arrows of the electric wave of greater amplitude.

b. A cross-sectional view of a magnetic-plane, waveguide tee junction is shown in figure 17. Here the electric intensity is assumed to be perpendicular to the paper, and the arrows are shown by dots as coming from the paper toward the reader. This situation occurs when a wavefront approaches the junction from the H arm. Since the geometry shows here that the polarity of the magnetic lines of force does not reverse, the waves leaving arms 1 and 2 are equal in amplitude and phase. The amplitude of equal-phase wavefronts entering arms 1 and 2 is added in the H arm. However, opposite-phase wavefronts entering arms 1 and 2 will cancel each other in the H arm when they have the same amplitude. If their amplitudes are not the same, the amplitude difference of the greater wavefront will appear at the H arm.

c. In the conventional hybrid tee described, all four arms leave the junction at right angles to each other. A modified version known as the forked hybrid tee is shown in figure 18. Arms 1 and 2 have been bent forward to point in the same direction. This design characteristic permits connection of arms 1 and 2 to radiating horns arranged in a square. The performance of the forked hybrid tee is the same as that of the conventional hybrid tee.

18. R-F TRANSMISSION

a. When the magnetron fires the r-f energy carried by the waveguide assembly is divided equally in phase and amplitude among the four horns of the antenna. Refer to figure 11. Intersections T1 and T4 represent conventional tee junctions. Intersections T2 and T3 represent forked hybrid tee junctions. The complete monopulse hybrid tee assembly is shown in figure 4-1 in a simplified form.

b. The r-f energy divides equally in amplitude between arms 1 and 2 of T1 with the energy leaving through arm 1 180° out of phase with the energy leaving through arm 2. This energy arrives at T2 and T3 through their respective H arms.

c. At T2 the energy divides equally in amplitude between arms 1 and 2 because this division is in the magnetic plane of the tee, the energy in arm 1 is
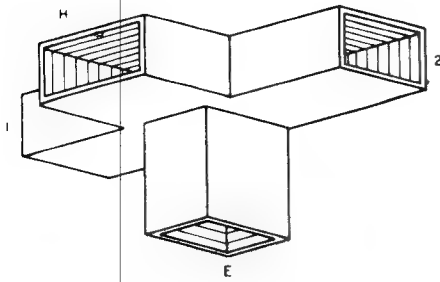
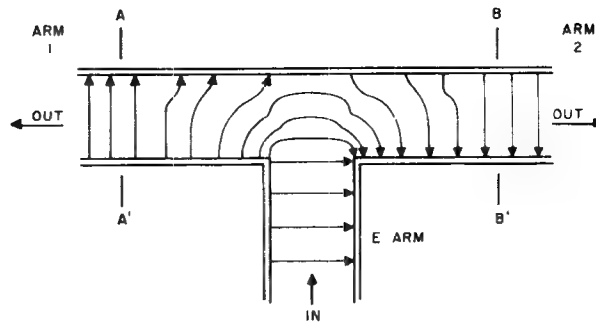Figure 14. Conventional wave-guide hybrid tee.

Figure 15. Spreading of a wavefront into an electric-plane junction from E arm.
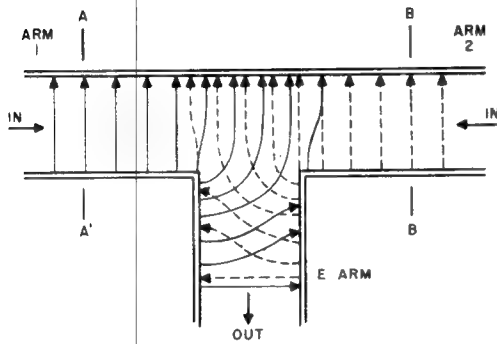

Figure 16. Spreading of a wave-front into an electric-plane junction from arms 1 and 2.
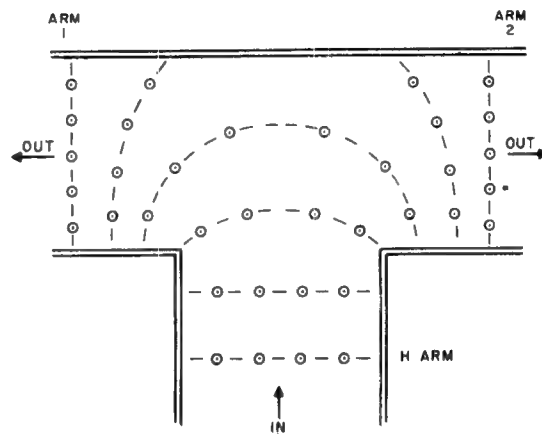

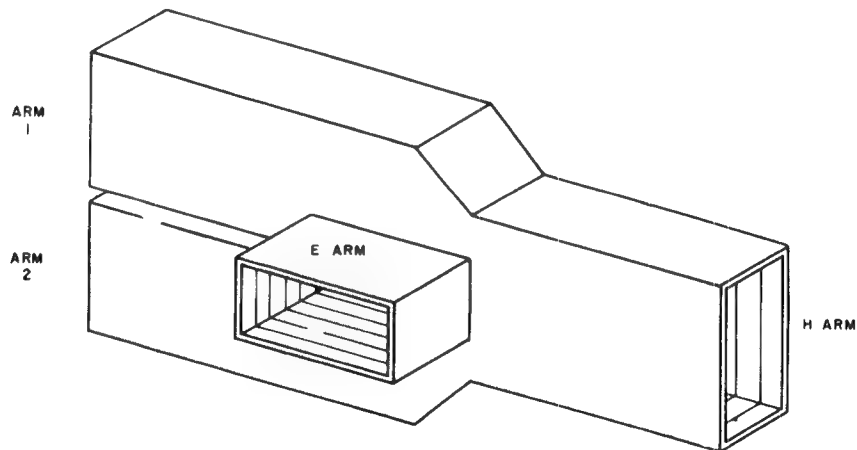Figure 17. Spreading of a wavefront into a magnetic-plane junction from H arm.

Figure 18. Forked hybrid tee.

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in phase with that in arm 2. Arms 1 and 2 of T2 lead directly to the upper left and right radiating horns. Likewise the energy entering T3 through its H arm divides equally in amplitude and leaves in phase through arms 1 and 2. These arms of T3 lead directly to the lower left and right radiating horns. Due to right and left 90° twist in the waveguide it can be seen that an in-phase relationship exists among the energies in all four radiating horns.

19. METAL LENS ANTENNA (figs 19, 20, and 21)

a. Track lens assembly GS-15645 consists of a number of parallel plates of varying widths mounted vertically so their edges face the mouths of the four horns. The theory of the lens antenna makes use of the optical properties of short-length radio waves. Light waves travel at a lower velocity through a conductive medium such as glass than through free space (fig 19). Since light rays at the edge of a convex glass lens travel through less glass thickness than those at the center, they emerge ahead of the center rays with respect to phase. A wavefront, connecting points of the same phase, thus appears concaved in the case of parallel light rays passing through a convex lens. This change in the contour of the wavefront thus concentrates a light beam at a focal point.

b. In the case of short radio waves, the parallel plates of the antenna lens act as waveguides (fig 20). When short radio waves travel through a waveguide their velocity is increased. Hence, unlike glass, the multiple waveguides in the lens increase rather than decrease the phase velocity of the waves. The antenna lens must therefore be made concave rather than convex so that the energy at the edges passes through a longer waveguide section and thus has a greater increase in phase velocity than the energy at the center of the lens. This produces the same concaved wavefront as did the convex glass lens with light waves, thus causing a focusing action. This theory has been stated for parallel rays being focused by a lens. The reverse action also takes place when waves originating at the focal point of the lens emerge from the lens as a beam of parallel rays.

c. The distance between the plates of the lens is between a half and a full wavelength of the radiating energy. The exact value of the length depends upon the required index of refraction (ratio of phase velocity of the energy in the lens to that in free space). The vertical plates are parallel to the electric vector of the electric-magnetic radio wave. The thickness of the supports separating the plates is small compared to the operating wavelengths, the separation of these supports is large compared to the wavelengths; thus the supports have negligible effect on the lens operation.

d. The side of the lens facing the radiating horns is so contoured that focusing occurs in all planes. The outer face of the lens is flat. This lens arrangement effectively produces a pencil beam of radiated energy having a

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beamwidth of 1.2 degrees at the half-power points. The four radiating horns are provided with a 0.25-inch lateral adjustment so that they may be located exactly at the focal point of the lens. This is a field adjustment.

e. To keep lens thickness at a minimum, the process of stepping is employed in which the lens profile is equivalent to a phase advance of one wavelength. Figure 21 shows a cross-sectional view of the stepping contour of a lens. The field pattern within a waveguide repeats itself at a regular wavelength interval along the waveguide. This spacing between points having the same phase advance is known as the guide wavelength. If a section of the lens, for example, required a piece of waveguide 3.8 guide wavelengths long, a waveguide 0.8 guide wavelengths long could be substituted, since the wavefront traveling through that section would still arrive at the focal point in phase with the rest of the wave. Therefore, no phase distortion occurs as a result of this stepping. The advantages of the stepping process are thus two-fold; first, the over-all thickness and thus the physical size of the antenna is reduced, and second, the radio waves are not as affected by inaccuracies in plate spacing and thickness because the waves pass through less of the frequency sensitive medium of the lens plates.

f. The lens antenna offers three advantages which the parabolic reflector antenna does not: the radiating horns are not in the way of the antenna beam because they are behind rather than in front of the antenna; the mean surface of the lens can suffer more warping without affecting the antenna beam, although the lens thickness and plate spacing must be moderately accurate; the lens antenna, despite its large size, is lighter and easier to construct and assemble than the continuous surface parabolic reflector antenna. The diameter of the track lens assembly GS-15645 is 6 feet. The focal length of the lens is 64 inches.

g. The track lens is covered on the outside (plate side) with a honeycombed plastic substance which provides better matching of the metal antenna to the air. This helps prevent reflections from the antenna itself. This lens coating also helps the antenna in a mechanical way by reducing the effect of the wind on azimuth tracking.

20. R-F RECEPTION

a. General (fig 23). The r-f pulse reflected by the target impinges on the lens area at a point whose distance to each of the four radiating horns determines the amount, by phase and amplitude comparison, of pulse energy applied to each of the four radiating horns. The radiating horns will then convey their signals, or respective portions of pulse energy, to the hybrid assembly where the mixing process takes place. Three signals are delivered by hybrid tee assemblies to the receiving system; namely, sum, azimuth difference, and elevation difference.
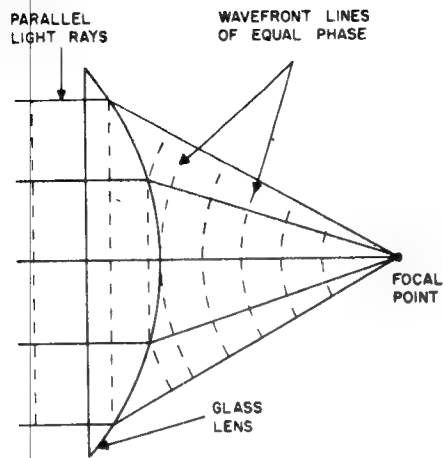
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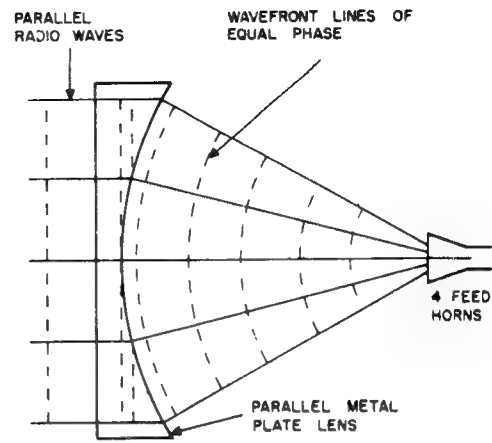Figure 19. Focusing of light waves with convex glass lens.

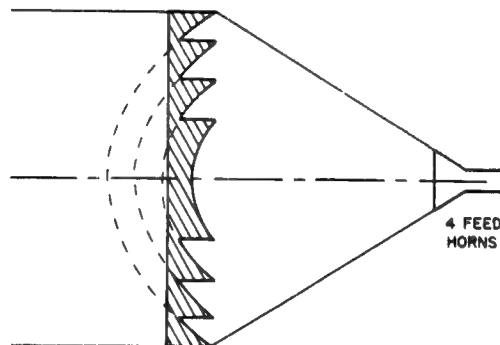Figure 20. Focusing of short radio waves with concave metal lens.

Figure 21. Cross-sectional view of metal lens illustrating stepping.

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b. No pointing error (fig 23). When the antenna is pointing on target the reflected signals will impinge upon the lens area at a point along the line AB. In this case, the right and left horns of the upper pair will receive the same amount of energy, and with the antenna being on target, the reflected signal will impinge upon the lens area at a point along line CD. This will assure that the lower pair of horns will receive the same amount of energy as the upper pair of horns (fig 22). At T2 and T3 the in-phase signals will cancel each other in the E arms. This precludes any energy being presented to T4 with a resultant azimuth error signal. Junction T1 is the point of separation of sum and elevation error signals. Energy is directed along the H arms of T2 and T3 toward junction T1. The energy arriving from T3 and T2 are 180° out of phase with each other, but equal in amplitude. Therefore, the two equal but opposite r-f energy signals will cancel in the H arms, with the resultant cancellation of any elevation error signal. However, at junction T1, the two r-f signals from the H arms of T2 and T3, due to hybrid action, are in phase and thus add in the E arm. This brings the entire energy flow from the four radiating horns into the E arm of T1, thus providing the sum r-f signal.

c. Pointing error (elevation and azimuth).

(1) In figure 23 assume that the target is located at the circumference of the circle at the arrow head denoting ANTENNA LENS. This places the antenna pointing off target in both azimuth and elevation. The target will reflect the strongest signal (———▶) the lower left horn, a slightly weaker signal (—— —▶) to the lower right horn, an even weaker signal (—▶——▶) to the upper left horn, and the weakest signal (—▶ —▶) to the upper right horn. Here, then, the lower left horn is furtherest away from the target and receives the greatest amount of reflected energy, because of the antenna action.

(2) The strong signal enters through the lower left feedhorn into arm 1 to T3. Simultaneously the weaker signal through the lower right horn enters arm 2 to T3. The energy flow is fed into the H arm of T3 in phase. However the two signals are 180° out of phase in the E arm, but of unequal amplitude. This difference of energy is fed through the E arm of T3 toward T4 and is indicated by the single arrow solid vector, the dominant energy flow.

(3) The weak r-f signal enters through the upper left horn into arm 1 of T2. Simultaneously the weakest signal enters through the upper right horn into arm 2 of T2. The energy flow is directed into the H arm of T2. However the signals are again 180° out of phase in this E arm. The energy directed into the E arm of T2 is the dominant signal of the two upper feedhorns and is denoted by the double arrow solid vector.

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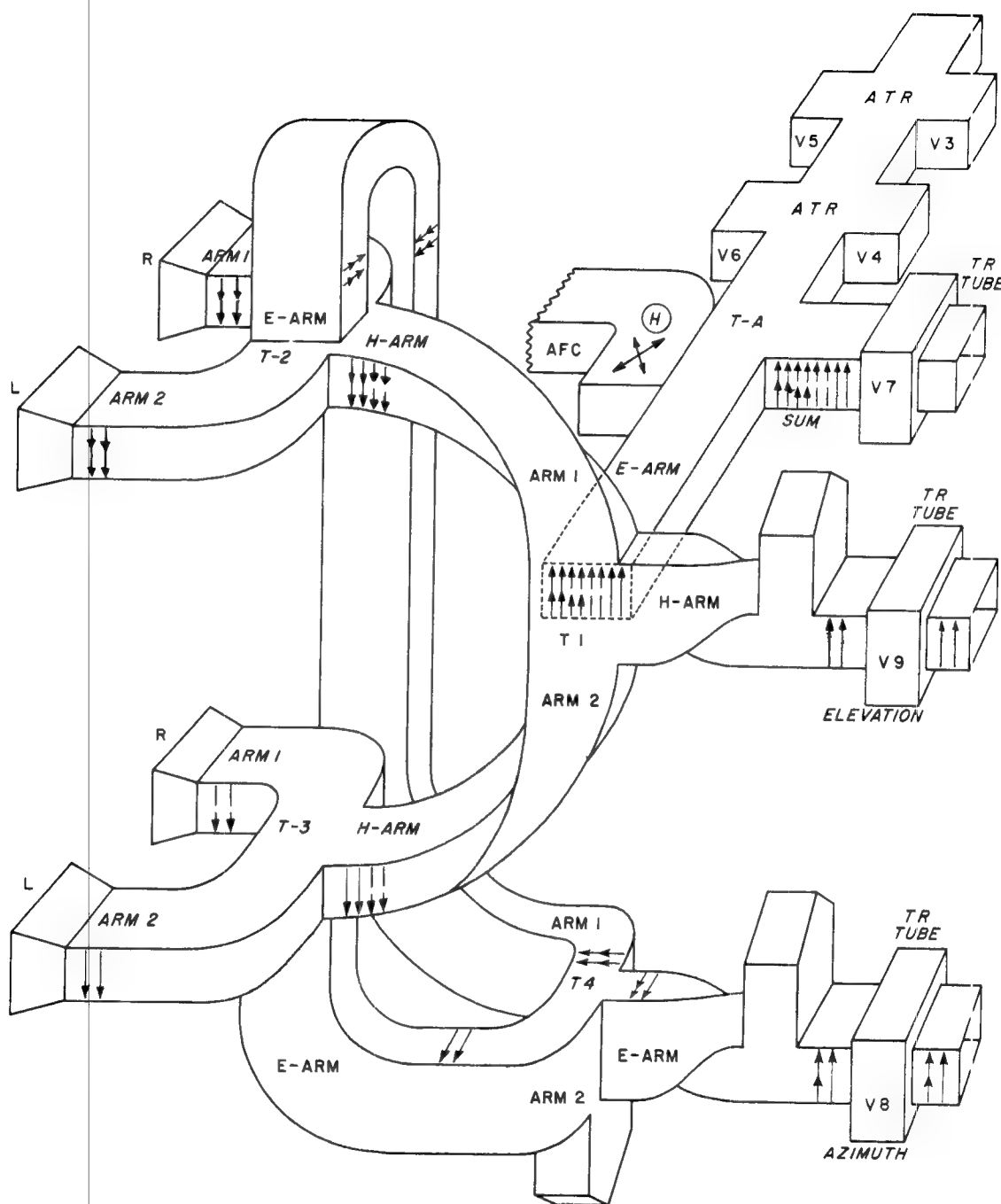
Figure 22. R-F transmission in monopulse radar.

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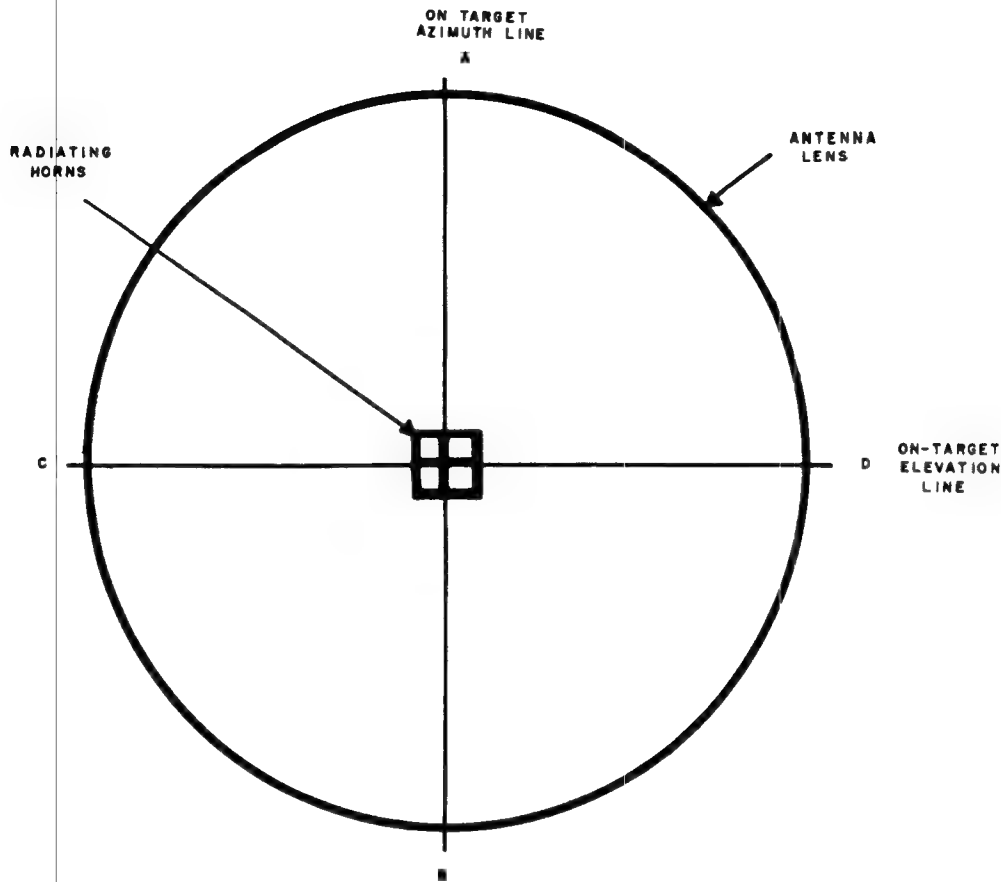
Figure 23. On-target comparison.

(4) The two dominant signals forwarded to T4, explained in the cases covered in (2) and (3) above, are in phase in the H arm of T4. The resultant r-f signal, indicated by a single and a double arrow solid vector, is the azimuth error signal.

(5) Returning to the r-f energy received at tee junctions T2 and T3, the received energy is in phase when it enters the H arms of the respective tee junctions and is therefore additive. This r-f energy is transmitted down the waveguide to junction T1. The energy, in phase in the E arm of junction T1, is transmitted to junction TA. However, the energy in the H arm is out of phase and tends to cancel. The dominating energy flow, indicated by a single arrow solid vector, results in an elevation error signal.

(6) As previously stated the r-f energy received at tee junctions T2 and T3 is in phase when it enters the H arm of the respective tee junctions and

therefore adds. The two paths of r-f energy are 180° out of phase with each other in the H arm of T1, but due to normal hybrid tee junction action they are in phase upon entering the E arm of T1. Thus the energy from both T2 and T3 add in the E arm of T1. This energy is transmitted down the waveguide to junction TA. The entire energy is then transmitted to the sum channel. The four ATR tubes prevent the received energy from entering the magnetron. Thus the sum signal is established and is used in the receiving system as a reference signal in the angle error detector.

(7) Should the antenna be pointing off target in the opposite direction, the same action described above would occur except that the dominating energy would be received at the upper right feedhorn. This would cause the corresponding azimuth and elevation error signals to be of a different phase in respect to the sum signal in the angle error detectors.

21. TR TUBE KEEP-ALIVE VOLTAGE

a. Purpose (fig 13). Keep-alive voltage provides quick ionization to TR tubes V7, V8, and V9 upon the arrival of the transmitted pulse. A few ions in the vicinity of the arcing cones of each tube are desirable. For this purpose a small discharge is maintained at all times in the interior of one of the cones of the tube. This discharge is produced by a third electrode in the TR tube, to which a keep-alive voltage of 200 to 300 volts is applied.

b. Power supply. Refer to figure 3-8. A separate power supply to furnish the -1,000 volts for keep-alive voltages is located in equipment panel A in target track r-f unit GS-15598. When the target-tracking radar is first turned on, 120- volt a-c power is applied to the primary of transformer T2 in equipment panel A. A source of 1,000 volts is developed across the secondary winding (5-6) of transformer T2. Thyratron rectifier tube (JAN5517) is connected to terminal 6 of transformer T2. This tube is connected as a half-wave rectifier with resistive-capacitive filtering in the plate circuit, supplying -1,000 volts with respect to ground. Resistors R23 and R26 and capacitors C6 and C7 form the filtering network. Resistors R24 and R25 are used as bleeders to discharge the filter capacitors with the power off. Switch E7 outside the equipment panel is a heavy spring contact on this equipment panel which connects the -1,000 volts to the waveguide assembly. The -1,000 volts is applied through resistors R33, R34, and R35 to the keep-alive electrodes of TR tubes V7, V8, and V9, respectively. A small electron flow from the keep-alive electrode to the walls of the cone produces the desired amount of continuous ionization with a measurable 200 to 300 keep-alive voltage.

22. ANTENNA BLOWER ASSEMBLY

See figure 3-8. To reduce the possibility of condensation in waveguide assembly GS-15600 of the missile and target r-f units, a blower assembly has

been added. This blower assembly consists of a phase shift capacitor C17, an air filter, a blower motor B4, and a flexible air line. Blower motor B4 is driven from the 120-volt, phase A power source. Air is pulled through the filter by the blower fan and forced into the flexible air line from which it passes to the antenna horn and in through the waveguide.

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CHAPTER 5

TTR RECEIVING SYSTEM

23. OVER-ALL SYSTEM

The purpose of the receiving system is to convert the r-f pulses applied from the r-f and antenna system into·intermediate frequency signals, to amplify these signals to a usable level, and to detect the signals. R-F pulses from four channels are applied to the receiving system: the sum channel pulses, the azimuth error channel pulses, the elevation error channel pulses, and the AFC channel pulses. By means of a local oscillator in the balanced converter, these four, r-f pulses are converted into 60-mc, i-f signals. The converter AFC signal is applied to an AFC unit whose output is used to control the frequency of the beating oscillator. The sum, elevation error, and azimuth error i-f signals are amplified and detected. The detected sum signal is applied in the form of video pulses to the ranging system, and the azimuth and elevation converted signals are applied in the form of d-c error voltages to the antenna positioning system. The gain of the amplified i-f signals is controlled automatically by an automatic gain control unit.

a. Components. The receiving system consists of track r-f converter assembly GS-15601; automatic frequency control (target) D-153964; two track azimuth and elevation i-f preamplifier units GS-15602; track sum i-f preamplifier GS-15691; and a section of slipring assembly GS-15536. All these units are located on the target antenna trailer. The remaining part of this system is located in the radar control trailer and consists of three track i-f amplifier units GS-15589, automatic gain control unit GS-15590, video and phase unit GS-15677, two angle error detector units GS-15718, and two error pulse rectifier units GS-15717.

b. Block diagram discussion. Figure 6-1 is the block diagram of the receiving system. The balanced converter, which includes the local reflex oscillator, transforms the X-band frequency r-f signals into 60-mc, i-f signals. The automatic frequency control unit maintains these signals at 60-mc frequency output. The 60-mc, i-f signals are preamplified to sufficient amplitude to reach the i-f main amplifiers through sliprings and 250 feet of cable. The precision-made sliprings transfer the three 60-mc, i-f signals from the rotating part to the stationary part of the antenna. These signals travel to the three i-f main amplifiers for further amplification from the antenna trailer to the radar control trailer through specially constructed triple coaxial cable. The long distance each of the three signals has to travel from the input to the i-f preamplifiers to the output of the i-f main amplifiers causes approximately 15,000° of phase shift.

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Since, the amount of phase shift is not the same in each channel, however, tracking error results which is detrimental to the monopulse system based on phase as well as on amplitude variations. To make each channel have as close as possible the same phase shift for zero error, the video and phase unit provides an adjustable phase shifting of ±80° in each of the three channels. The three channels are connected to their respective error detectors. Since the sum channel is concerned primarily with range, it supplies the video signals to the range error detector. However, to obtain relative direction of an error voltage in elevation and/or azimuth, the two signals are mixed with the sum signal in the two angle error detectors. The output of each angle error detector is in the form of video pulses. These are applied to their respective error pulse rectifiers which convert these error pulses into proportional d-c error voltages. The azimuth error pulse rectifier produces a d-c voltage which represents an error in the relative azimuth position of the antenna; the elevation error pulse rectifier produces a d-c error voltage in the relative elevation position. These d-c error voltages are applied to the antenna positioning system. The AGC unit keeps the output of the sum i-f main amplifier at a constant amplitude, regardless of amplitude variations of the input signal. The gain of the other two main amplifiers is separately controlled with a proportional voltage. The AGC unit is controlled by the strength of the sum video signal applied from the video and phase unit. The automatic gain control unit is gated; that is, it produces an output voltage only for the received pulses of the target being tracked. Gating of the signals is necessary to prevent the target radar main bang, other radar transmitted signals, and circuit noises from affecting the tracking of a single target. The i-f signal level monitoring and the gain information of the i-f main amplifiers are provided for calibrating, testing, and alining the units involved. When the target-tracking radar is used as a missile-tracking radar, facilities are provided in the video and phase unit to supply automatic frequency control signals to the missile AFC unit.

   c. Converter assembly (fig 24). Practical circuits for amplifying microwave signals do not yet exist, therefore it is not possible to amplify weak received signals in the form in which they appear at the antenna. Instead, the frequency of the received signals must first be reduced to a value that will permit amplication by conventional i-f stages. This reduction of frequency is the function of the converter. This unit system consists of a system of magic-T junctions where the various inputs are mixed together to produce the required i-f outputs. It is located in the r-f unit behind the waveguide assembly. The mechanical structure of the converter is shown in figure 6-2.

   (1) Inputs. There are five inputs to the converter assembly, four from the waveguide assembly and one from the local oscillator. The first input to be considered is from the local oscillator which is fed into the balanced converter and mixes in the converter with the other inputs to provide i-f outputs from the converter which have a frequency of 60

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megacycles per second. This intermediate frequency is maintained by
mixing a sample of the transmitted pulse in the AFC mixer with the
local oscillator signal. This transmitted pulse sample is taken from a
directional coupler in the waveguide assembly. If mixing of these two
signals produces a signal other than 60 megacycles, the AFC circuits
will cause the local oscillator signal to vary in frequency until the
mixing action produces the proper i-f signal. The third input is from
the waveguide assembly and is the sum signal. It is the sum of all the
r-f signals reflected from the target to the feedhorns. It is mixed in
the sum mixer with the local oscillator signal to produce a sum i-f
signal which is later detected and used in monopulse radar for video
display and automatic range tracking. The fourth inpus is the elevation
difference signal. It is an r-f signal from the waveguide assembly which
carries the elevation error information as reflected to the feedhorns from
the target. It is mixed in the elevation mixer with the local oscillator
frequency and produces the elevation i-f signal which is later detected
and used for elevation video display and elevation automatic tracking.
The fifth input is the azimuth difference signal. Its action are identical
to the elevation difference signal except for the information carried.
The elevation and azimuth difference signals are theoretically reduced
to zero when there is no pointing error.

(2) Outputs. The four outputs from the converter are all 60-mc, i-f sig-
nals. The first is the AFC signal which goes to the sum i-f preampli-
fier. The three other outputs are the sum i-f signal and the azimuth
and elevation i-f error signals which go to their respective i-f pre-
amplifier.

(3) Antijamming. The physical construction of the converter is such that it
will eliminate two signals entering it from the same source that differ by
60 megacycles; i.e., local oscillator noise; or, from the antenna, an
echo signal and a c-w jamming signal; or two jamming signals in the
frequency range of 8,500 to 9,600 megacycles. This is an important
antijamming feature of the target-tracking radar.

(4) Waveguide shutters. The crystals utilized by the mixing circuits could
be seriously damaged or ruined if high-power r-f energy were allowed
to enter the mixing circuits. TR switching provides this protection
during actual operation of the radar, but when the radar is deactivated
the TR tubes do not function. The crystals must then be protected from
strong pulses from any other nearby radars which may be in operation.
This protection is furnished by four waveguide shutters, flat strips of
metal, located between the three TR tubes and their crystals and between
the AFC directional coupler and the AFC crystals. The shutters, held
in place under spring tension when the transmitter is not operational,
are withdrawn after the magnetron warmup period by the same relay

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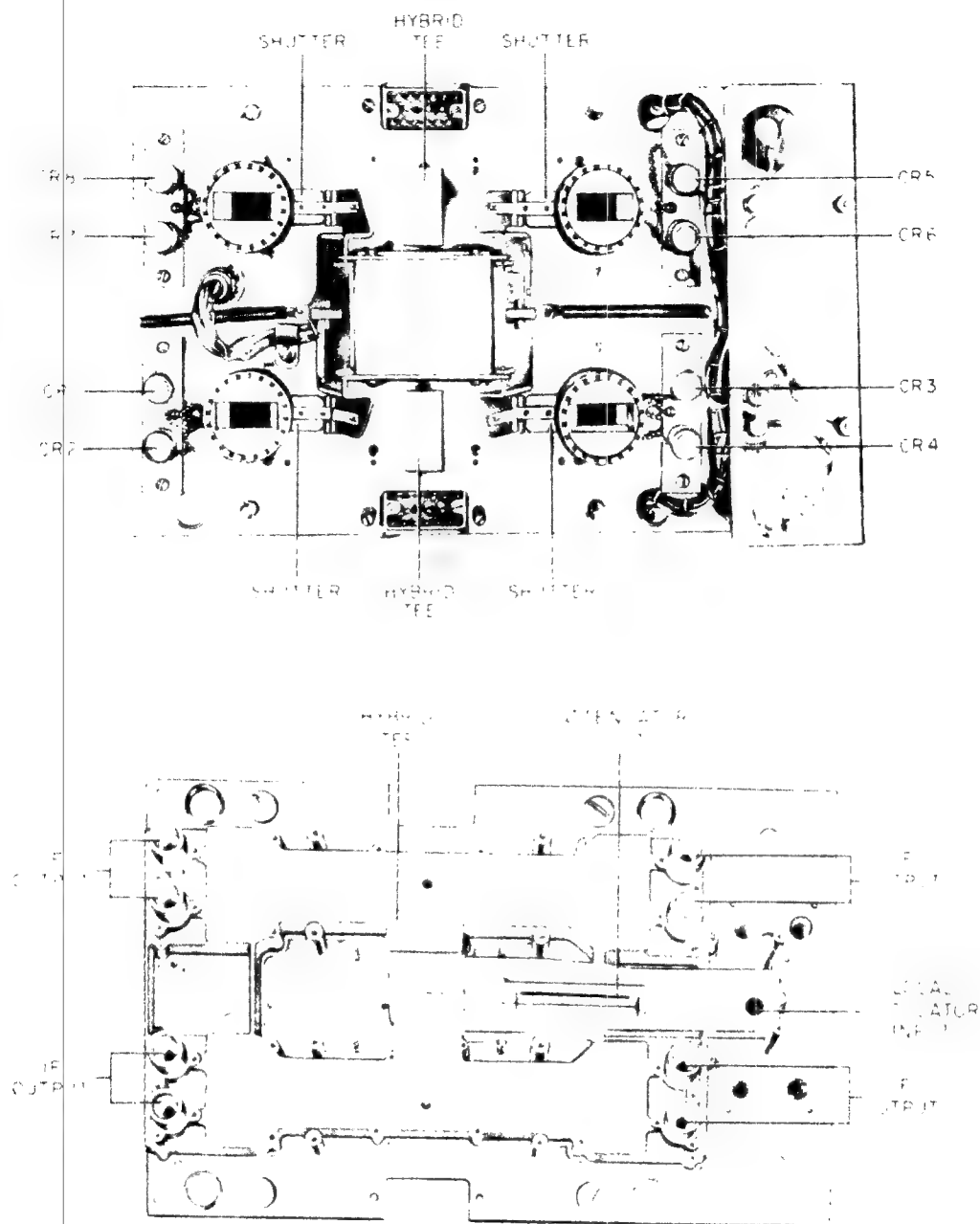
Figure 24. Converter assembly, waveguide structure.

that reduces magnetron heater current when the magnetron begins to conduct. The shutters are replaced in the waveguides when the radar transmitter is deactivated.
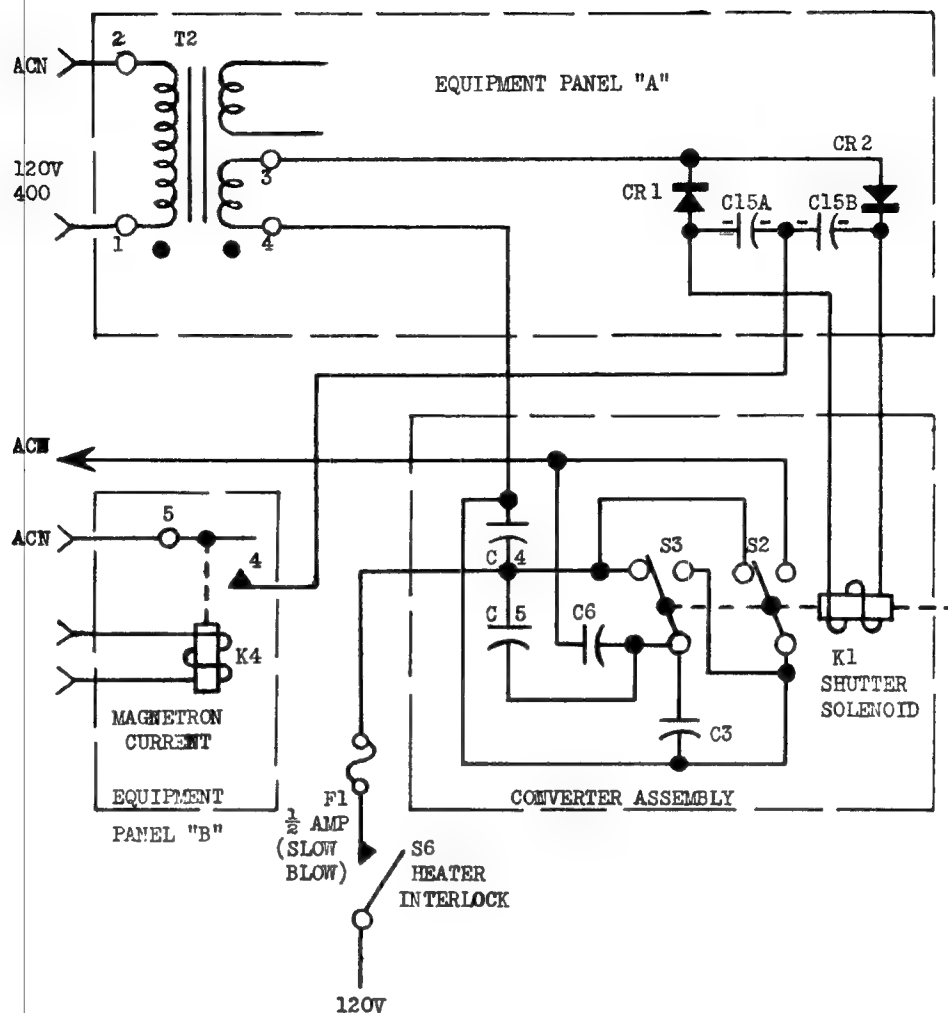
Figure 25. Shutter circuit, simplified schematic.

d. I-F preamplifiers. There are three i-f preamplifiers in the target radar, all located on the converter assembly in the r-f unit. These i-f preamplifiers sufficiently amplify the signals before transmission over cables to insure minimum noise pickup. The elevation and azimuth error signal channels have identical i-f preamplifier units , while the sum i-f preamplifier differs from these two only in that it has the necessary components for zero calibration of the range system and an AFC channel. The i-f preamplifiers are necessary because the signal must travel over a 250-foot cable before it reaches the radar control trailer and the amount of noise that would be picked up on the way would make

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the signal unusable. Since most of the noise is picked up in the first stage of i-f amplification, these preamplifiers use grounded-grid triodes to minimize this noise pickup and amplify the signals to a level necessary to overcome the loss in the cable.

(1) Inputs. The inputs to the i-f preamplifiers are the 60-mc, i-f (AFC, sum, azimuth, and elevation) signals from the converter assembly. The sum i-f signal from the sum mixer and the AFC i-f signal from the AFC mixer are transmitted to the sum i-f preamplifier. The elevation i-f signal from the elevation mixer is sent to the elevation i-f preamplifier, the azimuth i-f signal from the azimuth mixer, to the azimuth i-f preamplifier.

(2) Outputs. The outputs of the i-f preamplifiers are the amplified 60-mc, i-f signals which are transmitted over cables to the i-f main amplifiers in the radar control trailer and the 60-mc, i-f AFC signal which is sent to the target AFC unit. The sum i-f signal and the AFC signal, not altered in any way, are sent to the AFC circuits. The elevation signal is from the elevation i-f preamplifier; the azimuth i-f signal is from the azimuth i-f preamplifier.

e. **I-F main amplifiers.** The receiving system of the target-tracking radar employs three identical i-f main amplifiers; one for the sum i-f signal, one for the elevation i-f error signal, and one for the azimuth i-f error signal. Most of the gain of the receiving system is produced in these amplifiers. The i-f main amplifiers are located on the back wall of the radar range and receiver cabinet in the radar control trailer.

(1) Requirements. Nike I system specifications impose certain requirements upon the i-f main amplifiers. To maintain a steep leading edge of the i-f pulses, necessary for accurate range determination, the bandpass of the amplifiers must be 10 mc. To obtain maximum receiver sensitivity coincident with maximum range, great care is exercised to maintain maximum signal-to-noise ratio. Considerations arising from the use of monopulse require that the phase shift of the amplifier be held to a minimum (less than 30°) consistent with desirable gain characteristics. The amplifier must quickly recover following strong signals which momentarily block the circuit. It must be easily serviced and require a minimum number of adjustments.

(2) Inputs. The i-f main amplifiers receive their 60-mc sum, azimuth, and elevation input signals from the i-f preamplifiers via the sliprings and a 250-foot coaxial cable which connects the antenna trailer with the radar control trailer. A weak input signal, voltage-wise, may be amplified well over a million times by the i-f main amplifiers. The

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output of the AGC unit is fed into each of the i-f main amplifiers and controls the gain of the amplifiers. Both manual and automatic gain control of the receiving system are accomplished by varying the voltage on the control grids of V2 through V6 of the i-f main amplifier stages.

(3) Outputs. The outputs of the i-f main amplifiers are the amplified 60-mc sum, azimuth, and elevation i-f pulses, about 2-volt peak to peak, which are applied to the video and phase unit. The i-f test pulse from the i-f test set unit is used to aline and check out the receiving system. These units and those units that follow that are part of the receiving system are located in the target side of the radar range and receiver cabinet.

f. Video and phase unit. The proper operation of the angle error detectors and a subsequent pointing of the antenna demand that there be a 90° phase relationship between the sum channel and the two error channels when on target. Approximately 15,000° of over-all phase shift occur in each of the three channels from the time the signal first enters the feedhorns until it emerges from the i-f main amplifiers. Special precautions are taken to obtain exactly the same over-all phase shift in each of the three channels. However, with the large total phase shift, it would be impossible to have exact phase alinement if some circuitry were not provided to cancel out the relative phase shift between channels. Adjustments are provided in the video and phase unit to correct any such phase shift. The video and phase unit is located on the near wall of the range and receiver cabinet.

(1) Secondary functions. In addition to providing the phase necessary to obtain alinement, this unit accomplishes several other functions. First, it detects a portion of the output of each of the channels and then stretches the pulse so that a d-c meter on the test panel may be used to read the level of the output of each channel. Second, it provides test points where the unstretched i-f pulses may be monitored with an oscilloscope. Third, it provides two stages of i-f amplification.

(2) Inputs. Inputs to the video and phase unit are: the sum, azimuth, and elevation i-f signals from the three i-f main amplifiers; the 0.4μsec, positive 15-volt receiver gate from the range error detector, used to gate the monitoring circuits.

(3) Outputs. Outputs from the video and phase unit are: the sum i-f signal, which is sent to the angle detectors; the elevation i-f signal sent to the elevation angle detector; the azimuth i-f signal to the azimuth angle detector; a 0.3μsec, negative 1-volt video pulse from the sum channel to the AGC circuits; a 0.3μsec, positive 2-volt video pulse to the range error detector; a sum i-f signal to the missile AFC.

g. <u>Automatic frequency control unit (AFC).</u> During normal operation the frequency of the magnetron may be shifted several megacycles per second as a result of differing reflections from nearby objects. A change in the power level of the pulse applied to the cathode of the magnetron will also shift its frequency. Any change in magnetron or local oscillator frequency requires a correction in the local oscillator frequency to maintain the correct i-f beat frequency for optimum reception in the receiving system. The function of the AFC unit is to keep the local oscillator operating at 60 megacycles above the magnetron frequency and it continually changes the klystron frequency to accomplish this. The AFC system must operate rapidly enough to keep the klystron frequency from 55 to 65 megacycles above the magnetron frequency, a requirement imposed by the 10-mc bandpass of the i-f preamplifiers and i-f main amplifiers. The klystron operating frequency is controlled by varying the conduction of a tuning triode contained within the envelope of the klystron. The AFC unit also contains a search circuit to slowly sweep the frequency of the klystron up and down. This search system operates if the intermediate frequency gets out of the 10-mc bandpass. This search system also prevents the AFC from locking on at the point where the klystron frequency is 60 megacycles below the magnetron frequency. The sweep circuits will permit locking on only on the downward sweep of the klystron. (The input to the AFC unit is the i-f signal from the sum i-f preamplifier.) The signal is developed by mixing a sample of the transmitted pulse with the klystron output. The output of the AFC unit is a positive or a negative d-c voltage which is sent to the grid of the tuning triode to cause the klystron output to vary in the proper direction. From the klystron, the r-f signal is sent to the converter assembly where it is mixed with the other inputs to that unit. The AFC unit is located in the bottom right rear of the r-f unit.

h. <u>Local oscillator.</u> The local oscillator tube operates over the frequency range of the tracking transmitter to supply a c-w signal to the signal mixer and the AFC mixer. The output frequency is held 60 megacycles above the frequency of the transmitted and received signals, through the action of the AFC unit. In addition, the local oscillator, located in the left rear of the r-f unit, can be made to sweep through its frequency range to search for the proper operating frequency. The local oscillator tube is a 2K45 reflex klystron. This tube has a self-contained cavity whose effective capacitance is variable through the action of a tuning triode located within the same envelope. When the magnetron center frequency is varied by the radar operator, the same controls vary the repeller plate voltage of the local oscillator, carrying the output frequency along to be always centered at 60 megacycles above the transmitted frequency. For any given repeller plate voltage, the local oscillator may be made to vary in frequency over its entire frequency band by use of its tuning triode. As the effective capacitance of the cavity is varied, the resonant frequency also varies. The operation of the oscillator portion of the 2K45 klystron is based upon velocity modulation, in which the transit time of the electrons is controlled to reinforce the oscillations natural to the resonant cavity. Power is drawn off by means of a coaxial probe inserted into the resonant cavity through the base of the tube.

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Energy is transferred to the converter through a probe-to-waveguide coupling. The resonant frequency of the oscillator is changed by varying the effective capacitance of the cavity. This is done mechanically by varying the distance between the two cavity grids of the oscillator section. The mechanical motion originates in the tuning triode section of the tube.

    i.   Automatic gain control (AGC). An aircraft is a complex structure which when in flight and viewed from the ground presents a cross-sectional area that varies rapidly. It follows that when an aircraft is struck by pulses of electromagnetic energy from a tracking radar, the amplitude of the pulses reflected back to the radar also varies rapidly. This fading can occur in such a manner that there may be a large variation in the amplitude of successive pulses. With a lobing radar, these amplitude variations, if uncorrected, can introduce serious tracking errors. With a monopulse radar this is not the case since tracking data is derived from each pulse and is not based on a comparison of successive pulse amplitudes. Automatic gain control is still desirable in the receiving system of the target radar because it enables the angle error detectors to develop tracking voltages which are less noisy. This induces jitter in the servo loops. At the same time, the video error signal applied to the elevation and azimuth tracking indicators are improved, simplifying manual tracking by pip-canceling as employed in the target-tracking radar. To maintain the amplitude of the received pulses from a target at a more constant level, the i-f main amplifiers of the target-tracking radar are automatically gain-controlled. Gradual changes in the amplitude of the echo pulses resulting from changes in the range to the target are also compensated for by the AGC unit. When the target-tracking radar is exactly on target in the angle coordinates, the elevation and azimuth difference signals are reduced to zero. For this reason the gain of all three i-f main amplifiers must be controlled by a single AGC unit to which only one signal, the sum video signal, is applied. The sum video signal consists of negative video pulses detected by the video and phase unit and does not go to zero when the radar is on target. If the amplitude of the sum video signal exceeds a selected threshold level, the three outputs of the AGC unit are drived more negative and are applied as negative grid bias to the sum, elevation, and azimuth i-f amplifiers. The gain of the i-f amplifiers remains at a maximum as long as the amplitude of the sum video signal remains below the threshold level, but is decreased by an amount necessary to maintain the output of the i-f main amplifiers at a relatively constant amplitude. There are two inputs to the AGC unit. The first is the negative sum video pulse from the video and phase unit, and the second is the positive receiver gate which is used so that only signals from the ranged target will influence the AGC output. The output from the AGC unit is the negative grid bias voltage to the control grids of the second through the sixth stages of the i-f main amplifiers. There are provisions in the range and receiver cabinet for monitoring the AGC output by use of a meter located in that cabinet. This AGC unit is the fastest operating AGC unit of its type now in use. Though it may require several pulses to

produce a change in the AGC output, its control is fully effective from d-c to 30 cycles per second and still provides some degree of control at 135 cycles per second.

j.   <u>Angle error detectors</u>. Two angle error detectors are used in the receiving system, but since the two are identical only the azimuth unit will be discussed. The received azimuth error i-f signal and a portion of the received sum signal are applied to the angle error detector. This unit uses these signals to produce a video pulse signal which represents the direction and magnitude of the azimuth antenna pointing error. In ideal radar operation, if the antenna is off target in azimuth in one direction the azimuth error i-f signal will be 180° out of phase with the sum i-f signal. If the antenna is off target in azimuth, the amplitude of the output video pulse is proportional to the amount of azimuth error. If the antenna is pointing on target in azimuth, no azimuth error i-f signal is applied to this unit; this results in no output from the unit. As the target moves in one direction increasingly away from the antenna azimuth on-target position, the output of this unit will be negative video pulses of increasing amplitude. If the target moves in the opposite direction in the same manner, the output of the unit will be positive video pulses of increasing amplitude. The 3-microsecond expansion pulse from the ranging system is also applied to the angle error detector; this signal is used to gate the output video pulses, which insures that only the received i-f signals from the target being tracked are utilized. A portion of the output video pulses from the angle error detector is applied to the presentation system where it is used in pip-canceling on the azimuth indicator. The angle error detectors are located on the back wall of the radar range and receiver cabinet.

k.   <u>Error pulse rectifiers</u>. Two error pulse rectifiers are utilized in the target-tracking radar; one for the elevation channel and one for the azimuth channel. They are located on the inside right wall of the radar range and receiver cabinet. The error pulse rectifiers accept the angle error video pulses from the respective angle error detectors and convert the information contained in the pulses into d-c signals. These d-c signals are then applied to the antenna positioning system. When the antenna is pointing on target no angle error video pulses are applied to the error pulse rectifier and its output is a zero d-c signal with respect to ground. If the antenna is pointing off target such that positive video pulses are applied to the unit, a positive d-c signal will be its output; and for negative input pulses, its output will be a negative d-c signal. Since the amplitude of the d-c signals varies proportionaly with the peak amplitude of the video signals applied, they are proportional to the magnitude of the antenna pointing error. The error pulse rectifiers are gated by the 0.4-microsecond receiver gate from the ranging system. This gate permits only the error video from the target being tracked to affect the output of the unit.

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24. BALANCED CONVERTER (fig 26)

The balanced converter consists primarily of five separate sections; the local oscillator (reflex klystron), AFC converter, azimuth converter, sum converter, and elevation converter. The crystal current meter is not an integral part of the balanced converter but will be covered in the discussion of the balanced converter operation. Since all sections of the balanced converter sections operate the same way, only the azimuth converter will be discussed.
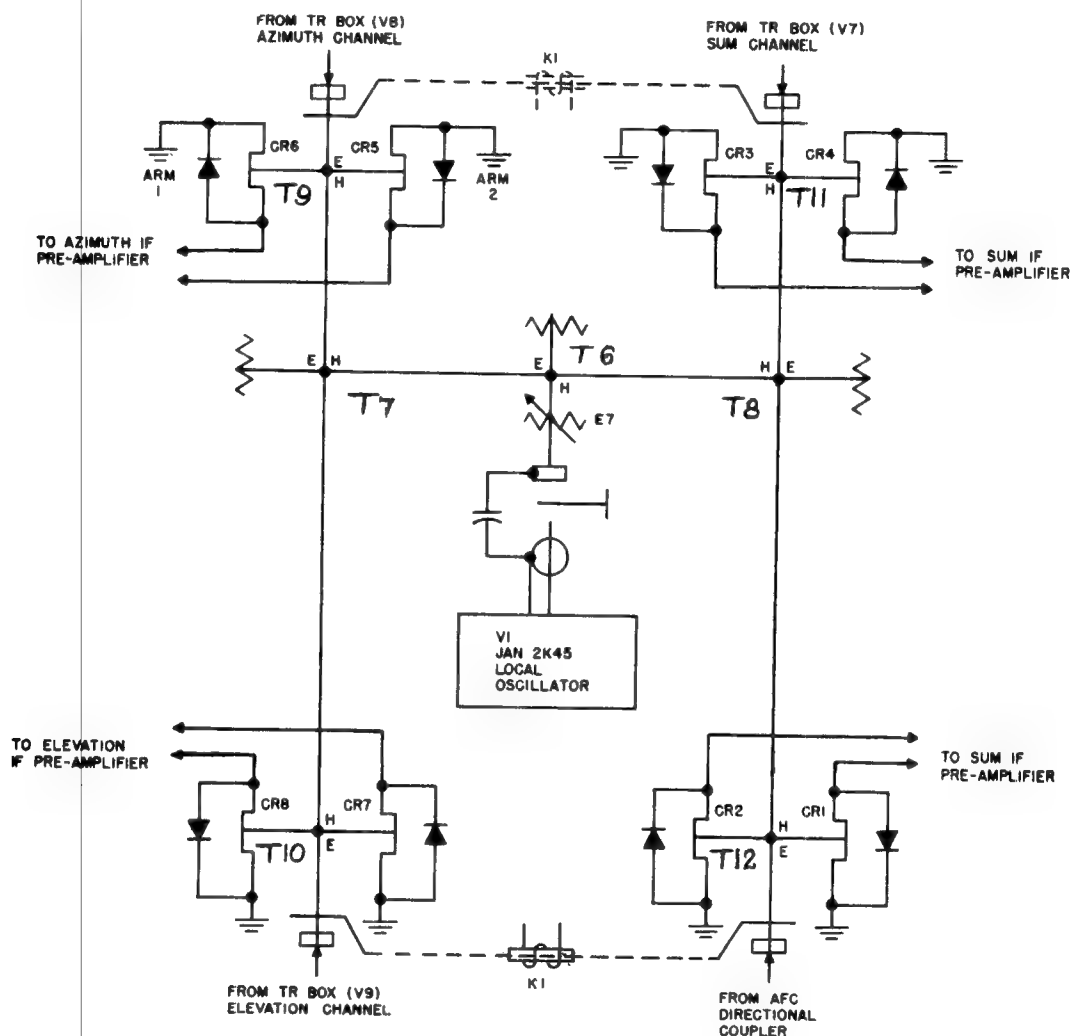

Figure 26. Converter assembly, simplified schematic.

a. Block diagram discussion (fig 6-2).

(1) Azimuth converter (fig 26). The azimuth deviation or error signal detected at the radiating horns is fed through a complex waveguide

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system to TR tube V8 where it is channeled to crystals (varistors) CR5 and CR6 with all frequencies but the 55- to 65-mc i-f signal canceled in the i-f preamplifiers. The output of the local oscillator is also fed to hybrid tee junction T9 and mixes with the received pulse where only the difference frequency obtained by mixing is used. The i-f pulse is fed to the azimuth preamplifier.

(2) Local oscillator (figs 26 and 27). The local oscillator is used as a beat frequency oscillator to produce a usable i-f signal in the receiving system. The frequency of the oscillator is controlled by the adjustment of the repeller plate voltage and also by a control voltage on the tuner grid. The frequency is variable, but is held at a constant 60 megacycles above the transmitter frequency.

b. Detail circuitry discussion. Refer to figure 26. The track r-f converter assembly GS-15601 consists of four waveguide shutters and their solenoid K1, the waveguide plumbing, mixing crystals CR1 through CR8 (trade-named varistors), at the local oscillator tube V1 with its temperature controlling system, and the crystal current metering circuit.

(1) Waveguide shutters (fig 24). The shutters are flat strips of metal inserted into the waveguide through slots. Three of the four insertion slots are located between the three TR tubes (V7, V8, V9) and the respective crystals. The fourth slot is in between the output of the AFC directional coupler and the AFC crystals. When the transmitter is not operating, normal protection by the TR tubes is not available, thus exposing the crystals to damage from energy directed into the antenna horns from nearby radars. As a protective measure, when the transmitter is not operating, the shutters are held under spring tension across the waveguides, completely blocking the flow of energy into the receiver. The shutter inserted into the AFC channel is not absolutely necessary since sufficient signal strength could not get through the directional coupler to damage the crystals. When the transmitter is operating and after the magnetron has warmed up, the shutters are withdrawn from the waveguides by a lever arrangement driven by shutter solenoid K1, thus preparing the receiver for normal operation. The shutter solenoid is a strong magnet which, when energized, pulls the metal armatures into the solenoid center. Each armature is connected to a lever plate by a connecting rod. Two waveguide shutters are connected to each connecting rod and they in turn are connected to a lever. Both plates are hinged so that when the solenoid is energized the armature pulls both plates toward the center of the solenoid, removing the attached waveguide shutters. The plates are spring-loaded so that upon deenergizing the shutter solenoid, the shutters are returned firmly in their slots. Figure 25 is a simplified schematic of the shutter control circuit. When the radar is first turned on, MAG
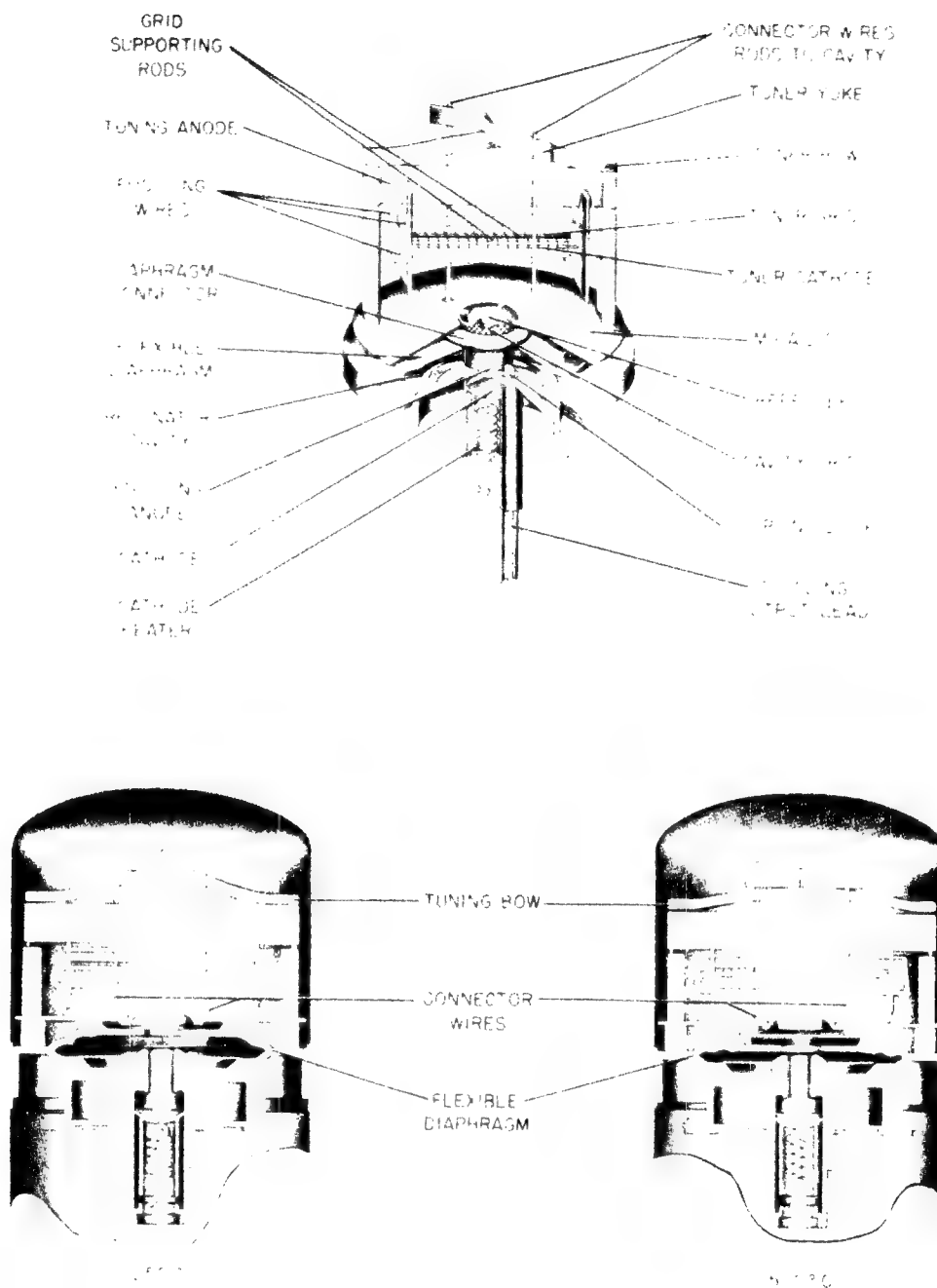
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Figure 27.   The 2K45 reflex oscillator.

CURRENT relay K4 in equipment panel B is deenergized since the magnetron is not drawing current. To winding 1-2 of transformer T2 (equipment panel A) is applied a-c neutral and 120-volt, 400-cycle relay (K) voltage. Single phase 120-volt power is supplied to one secondary of T2 (3-4 winding) for the shutter solenoid K1 on the converter assembly. As long as MAG CURRENT relay K4 is deenergized, however, winding 3-4 of transformer T2 is an open circuit. When MAG CURRENT relay K4 is energized, after the five-minute warmup of the of the magnetron, its contacts place a-c neutral at the junction of capacitors C15A and C15B. To terminal 4 of transformer T2 is connected 120-volt 400-cycle voltage through fuse F1 and the contacts of deenergized shutter solenoid K1. Therefore, across the secondary of T2 (3-4 winding) appears the sum of two voltages, 120 volts applied to the winding and 26 volts induced from the primary, making a total of 146 volts. This added voltage is used as an initial surge which develops enough power in the shutter solenoid to overcome the tension of the shutter springs. After the shutters are removed the solenoid remains energized by its induced voltage only. Diode crystals CR1 and CR2 and capacitors C15A and C15B form a full-wave rectifier circuit with voltages adding. When crystal CR2 conducts, capacitor C15B charges, and when CR1 conducts, capacitor C15A charges. Both capacitors charge with the polarities indicated in figure 25. With an effective voltage of 146 volts across the secondary of transformer T2, approximately 110 volts appears between the left plate of capacitor C15A and the right plate of C15B. (The capacitors do not charge up fully to the maximum voltage of the sine wave, only to about 80 volts each.) This voltage appears across shutter solenoid K1 which then energizes and removes the shutters. The total voltage across the relay is about 160 volts minimum. When shutter solenoid K1 energizes, its contacts, S2 and S3, apply a-c neutral to terminal 4 of transformer T2, at the same time removing 120 volts. So now with solenoid K1 energized, only the induced voltage (26 volts) from the primary of T2 appears across winding 3-4 of T2. The d-c voltage on solenoid K1 is at this time about 30 volts, sufficient to hold solenoid K1 energized.

(2) Converter waveguide plumbing. There are seven hybrid tees in this converter. They are appropriately joined to channel the energy from various sources in such phase and direction that mixing takes place in the eight crystals. From this mixing, the desired i-f output signals are obtained. The following signals are traced in figure 26:

(a) The received signal energy from the azimuth channel TR tube enters (shutters are open) hybrid tee T9 and divides equally into the tee's two arms. Each arm contains a mixing crystal (CR5 and CR6).

(b) The received signals from the sum and elevation channel TR tubes travel individual paths similar to the azimuth channel. Each sum

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and elevation signal appears through T11 and T10 respectively at the proper crystals (CR3, CR4, and CR7, CR8 respectively).

(c) The received signal energy from the AFC channel directional coupler enters through hybrid tee T12 and divides equally into the two arms, each containing a crystal (CR1 and CR2).

(d) The local oscillator signal comes from tube V1 into hybrid tee T6 and divides equally into two arms. Each arm has a hybrid tee (T7 and T8) which again divides the signal equally. Finally, the oscillator signal is divided equally into the eight arms containing the eight crystals.

Summing up the preceding flows of signal it can be seen that each of the eight crystals receive the same amount of oscillator signal. Also, each of the two azimuth crystals receive the same amount of azimuth difference signal, the two elevation crystals receive the same amount of elevation difference signal and likewise for the sum and AFC crystals. Across the waveguide are various attenuating devices which assure that the signals arriving at the crystals are at correct power levels for optimum operation. Attenuator E7 in the waveguide arm from the local oscillator is a resistance card whose position can be varied to control the flow of energy to the crystals.

(3) Signal conversion. Conventional hybrid tees as well as forked hybrid tees are incorporated in the balanced converter. (For a detailed discussion of hybrid tees, refer to chapter 4, paragraph 17.) The description of signal conversion which follows applies equally to each of the four receiving channels. Only the azimuth channel is discussed. Refer to figure 26. The azimuth error difference, returned r-f signals from TR tube V8, enters through the E arm of hybrid tee T9 containing CR5 and CR6. This means that the r-f signal is divided equally in amplitude between arms 1 and 2 of T9, but the signal in arm 1 is 180° out of phase with the signal in arm 2. The r-f signal from the local oscillator enters T9 through the H arm and divides equally in amplitude between arms 1 and 2 of T9. These two signal components are also in phase. Mixing of the returned r-f signal and the local oscillator r-f signal in arms 1 and 2 of T9 yields several signals of different frequencies (8,500 - 9,600, 8,560 - 9,660, 17,060 - 19,260, and 60 megacycles). The frequency concerned here is the difference frequency (60 mc) between the two r-f signals. Because the local oscillator frequency is continually variable and controlled, the difference in the two signals is an almost constant 60 megacycles. The input and output circuits of the two crystals are broadly banded with 60 megacycles as the center frequency. Therefore, the 60-megacycle difference frequency (or i-f) component is predominant in amplitude and is also applied to the crystals. The crystals are

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nonlinear devices that conduct current more readily in one direction than in the other. Therefore, the beat frequency envelope (i-f) of the two original radio frequencies appears at the output of the crystals. Since the received r-f signal component in arms 1 and 2 of the tee are 180° out of phase, it would appear that the outputs of crystals CR5 and CR6 would have this phase relationship. One crystal, however, is physically inserted oppositely to the other, thus reversing the phase relationship and causing the outputs of the crystals to be in phase. The two crystal outputs are coupled together through capacitors in the azimuth i-f pre-amplifier. These signals are connected from the gold-plated crystal holders containing the crystals through gold-plated tension springs on the preamplifier. In the preamplifier the two signals add. Similar explanation is applicable to the other channels of the converter. The balance converter eliminates unwanted i-f signals created when two signals enter the converter from the same source; i.e., local oscillator signal and local oscillator noise; an echo signal and a c-w anti-jamming signal from the antenna; or two jamming signals anywhere in the frequency range of 8,500 to 9,600 megacycles. Assume the existence of two jamming signals within the range 8,500 to 9,600 megacycles separated by 60 megacycles. One of these signals enters through the E plane arm and divides equally in magnitude, but the signal in one crystal arm is 180° out of phase with the signal in the other crystal arm. The second jamming signal divides similarly and the resulting 60-mc mixture in each crystal arm will be in phase with each other. The two unwanted 60-mc, i-f signals leave the crystals 180° out of phase with each other. When these two signals join the i-f preamplifier, they cancel. Similarly, two signals from the local oscillators separated by 60 megacycles (producing an unwanted i-f signal) enter the crystal arms of the tee with no phase shifts, resulting in 180° phase difference at the outputs of the crystals and cancellation in the i-f preamplifier. If a single crystal mixer were used in each channel, spurious echoes, range jamming, or high noise would appear in the output. The use of a double crystal mixer in the balanced converter, however, eliminates fake i-f signals from external sources. The AFC channel is less concerned with the noise problem than the other three channels because the only sources of AFC signals are from the magnetron and local oscillator.

(4) Local oscillator.

(a) Although figure 3-8 identifies reflex oscillator JAN 2K45 as a "beating" oscillator, it is referred to as a local oscillator in this text. The local oscillator operates over the frequency range 8,500 to 9,600 meg-acycle to supply beating signal energy to the balanced converter. The oscillator frequency is held 60 megacycles higher than the incoming signal frequency (actually the magnetron frequency) through the action of the AFC unit. In addition the local oscillator can sweep automatically

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over one-third its entire frequency range for a given repeller voltage to search for the received signal when the transmitter frequency or magnetron frequency suddenly shifts. Local oscillator JAN 2K45 has a self-contained cavity whose resonant frequency is variable through the action of a tuning triode, located within the same envelope (pins 1,2,3), and the repeller plate voltage. Figure 27 shows an internal view of the tube.

Portion

(b) The operation of the oscillator ~~position~~ of the tube is based on velocity modulation of the electron transit time. Electrons leaving the cathode toward the repeller pass through the two resonator grids, shocking the cavity into low-level oscillation at its resonant frequency. Alternating fields of the same frequency are set up between the resonator grids. Electrons passing between the grids when the field strength is zero undergo no change in velocity. Electrons passing through at other times are accelerated or decelerated in proportion to the direction and intensity of the field. In their travel toward the repeller, therefore, electrons tend to form bunches. By adjusting the repeller voltage to a suitable negative value with respect to the cathode, the electron bunches are caused to return through the resonator grids in such phase that the the cavity oscillation is reinforced. Power is drawn off for use in the converter through a coaxial probe inserted into the cavity from the base of the tube. Energy is transferred to the converter through a probe-to-waveguide coupling. The tuning triode has a specially constructed anode which expands as a result of heat developed by electron bombardment. The expansion is a function of the electron flow through the triode. The tuner bow attached to the ends of the anode translates the lateral expansion of the anode to vertical motion. This motion is applied through the tuner yoke and two rods to the flexible diaphragm forming the top of the reflex oscillator cavity. The rods also connect +320 volts to the second resonator grid, which acts as an accelerating electrode for the electron stream. As current through the tuning triode increases, the temperature of the anode rises, the cavity size and resonator grid spacing decreases, and the resonant frequency of the cavity decreases. Figure 27 illustrates this action.

(c) The combination of the tuning triode and an adjustment of repeller plate voltage enables the reflex oscillator to be tuned over a frequency range of 8,560 to 9,660 megacycles (fig 28). When the center frequency of the magnetron is varied, the repeller plate voltage is also varied. Potentiometer R17B in the magnetron tuning drive unit varies this repeller voltage through the required range. The potentiometer arm is driven by the magnetron tuning motor B2.

(d) The magnitude and spread of repeller voltage are established by the setting of potentiometers R20 and R18, which are part of a voltage

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divider across -250 volts in equipment panel A. Grid voltage for the tuning triode is obtained from AFC unit D-153964. When the local oscillator operates under automatic frequency control, this voltage is a d-c bias that changes rapidly in polarity whenever necessary to keep the oscillator tuned 60 megacycles higher than the magnetron. The AFC unit also applies a control voltage whose magnitude causes the local oscillator to search over one-third of the frequency range in order to regain the i-f signal. In this case, the voltage applied to the tuning triode is in the form of a square wave having a repetition rate of one cycle every 15 seconds. During the positive excursion of the square wave, maximum conduction takes place in the tuning triode causing gradual compression of the oscillator cavity to a minimum size. After $7\frac{1}{2}$ seconds the square wave reverses, cutting off the tuning triode. The tuning anode cools off and the cavity gradually returns to its former size. Further discussion of automatic frequency control is given in paragraph 27. Local oscillator tube V1 is enclosed in a metal shield to prevent oscillator radiation to other circuits. Radio frequencies are filtered from the leads to the tube by impedance networks Z1 through Z4. These chokes, shown on figure 28 as a block across the tube connections, are polyiron cylinders mounted concentrically with the incoming leads.

(e) The local oscillator V1 generates a great amount of heat with the tuner triode operation. The tube is covered with a metal shield for r-f shielding purposes. Therefore, to prevent the tube from being over heated, blower motor B3 directs a stream of air onto the tube for cooling. The blower is thermostatically controlled by thermostat S4 which keeps the temperature of V1 at approximately 110° C.

(5) Crystal current metering. Provisions are made to check converter operation by the metering of all eight crystals in the converter. Refer to figure 3-8. Crystal current meter M1 and crystal selector switch S1 are located on a metal plate directly attached to the back of the converter assembly. This metal plate also supports the three i-f preamplifiers. The crystal currents pass through the three i-f preamplifiers and these are filtered of all r-f variations. Four of the crystals (from four channels) are inserted into the waveguide in the opposite position from the other four crystals (in the four channels). Hence, four crystal currents will flow in one direction and the remaining four currents, in the opposite direction. Switch S1, in conjunction with resistors R6 through R13, is connected to XTAL CURRENT meter M1 in such a way that currents flowing from the crystals in opposite directions will be applied to the meter correctly. The AFC crystal current may also be metered in track test panel GS-15597 in the radar control trailer, however only crystal CR1 is metered in this test panel (fig 29). When switch S1 in the converter is in either OFF position, the flow of electrons from CR1 is as follows: Electrons leave pin 9 of plug P3 in the converter, pass through sliprings and a cable
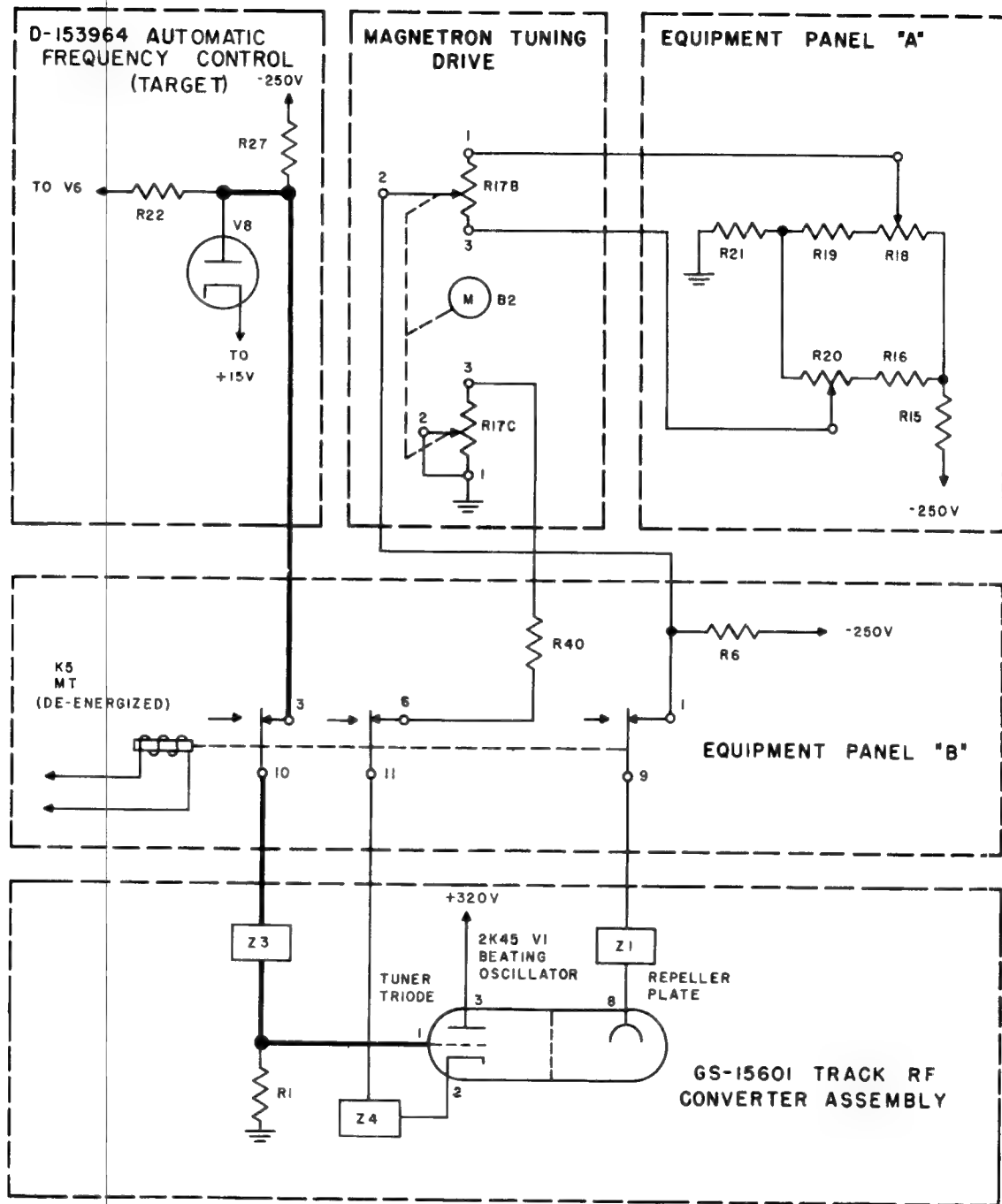
Figure 28.   Target-tracking radar local oscillator control.

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and enter the track test panel GS-15597 at pin 1 of P1. See figure 6-23.1. With TEST SELECTOR switch S1 (in this panel) in the XTAL position, the electron current is sent through meter M1 and resistor R1, which are in parallel. It then passes out pin 3 of P1 in the test panel and returns to the converter through pin 11 of P3. It then goes through resistors R6 and R7 and switch S1 to ground. When metering the CR1 current at the converter, meter M1 in the converter is placed in series with the path just described by the switching action of S1A and S1B. The crystal current in the radar control trailer is for the purpose of checking local oscillator operation.
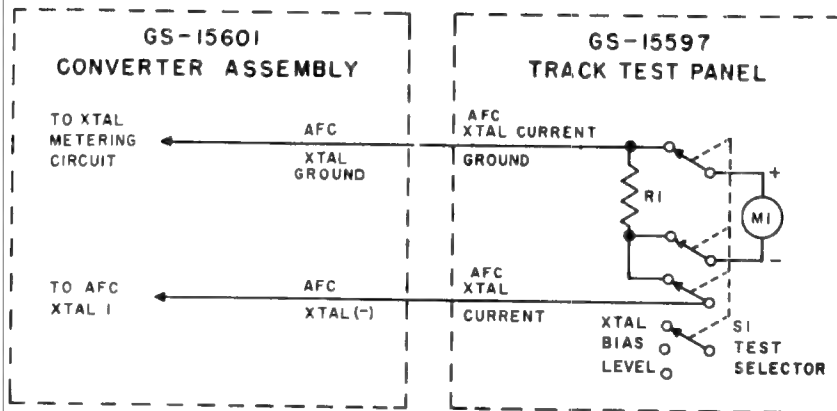
Figure 29. AFC crystal metering circuit, simplified schematic.

25. ERROR I-F PREAMPLIFIER (fig 6-9)

There are two error i-f preamplifiers used in the receiving system of the target-tracking radar, but since their operation is identical, only the operation for the azimuth i-f preamplifiers will be discussed. The i-f preamplifier consists of two noise limiting capacitors in the input circuit, a grounded-grid i-f amplifier, a conventional class A i-f amplifier, and two r-f filter networks. The output of the i-f amplifier is fed to the azimuth i-f main amplifier through the slipring and trunnion assembly.

a. Block diagram discussion.

(1) Grounded-grid i-f amplifier. Since most noises are picked up in the first stages of i-f amplification, the grounded-grid amplifier is used to reduce the noise in relation to the signal. The grounded-grid amplifier was developed as the answer to the problem of neutralization of r-f amplifiers at high frequencies. The grounded grid acts as a shield to reduce the plate-to-cathode feedback capacitance in the same manner that the screen grid isolates the plate and cathode in a pentode or tetrode. In the conventional circuit, the suppressor and screen grids of tetrodes and pentodes reduce the feedback capacitance to a point where neutralization is seldom necessary up to frequencies in the neighborhood of 100

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megacycles. At frequencies of several hundred megacycles the screen and suppressor lead inductance makes it impossible to hold these leads at zero r-f potential and oscillation is likely to occur. In this case grounded-grid amplifiers are used. Another characteristic of the grounded-grid amplifier is that the instantaneous r-f signal voltage developed between cathode and ground is in phase with the instantaneous r-f signal voltage developed between plate and cathode. These voltages appear in series across the load. Driving power requirements for a grounded-grid amplifier are greater than for the same tube used as a conventional amplifier because the driver supplies some of the output power appearing at the load of the grounded-grid stage. However, this added power, which may be up to 10 times the driving power requirements of a conventional amplifier, is not lost, but simply transferred to the output circuit of the grounded-grid amplifier.

(2) I-F amplifier. The i-f amplifier is a conventional amplifier which feeds the 60-mc, i-f pulses to the azimuth i-f main amplifier through the slipring and trunnion assembly.

(3) R-F filters. The two r-f filters are identical in components and purpose. They are pi-type filters that filter out the r-f signal leaving only the d-c voltage on the outputs to the crystal metering circuit.

b. Detail circuitry discussion. The main i-f amplifiers for the receiving system are located approximately 250 feet from the balanced converter in the antenna assembly. The three i-f signals are transferred through sliprings and coaxial cables from the antenna to the radar control trailer. In this transmission there occurs a signal loss of about 4 decibels. Since the largest output of the balanced converter is minute, i-f preamplifiers have to be used at the antenna. The azimuth i-f channel and elevation i-f channel each incorporate a track azimuth and elevation i-f preamplifier GS-15602. Since the same type unit is used in both channels, only one discussion is necessary.

(1) Refer to figure 6-10. Track azimuth and elevation i-f preamplifier GS-15602 is located in the target track r-f unit GS-15598 on the target track antenna. The preamplifier is attached directly to the converter assembly. The 60-mc output of the converter is applied to the preamplifier through a small spring under tension touching the crystal holders in the converter. The two stages, V1 and V2, are grounded-grid triodes which give an amplification of about 23 db but improve the signal-to-noise ratio. Heavy loading of the input and output circuits eliminates the natural tendency of triodes to break into oscillation and maintains a broadband response necessary for faithful reproduction of the input signal. Z1 at the input of V1 is a factory-adjustable inductance which is used to tune the input. The input is tuned slightly inductive

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for the purpose of obtaining a better signal-to-noise ratio. The two inputs come from the two crystals in the balanced converter and are coupled through capacitors C1 and C2 to the cathode of V1. Any spurious noise signals developed within the converter crystals, due to operation of the converter, are applied through capacitors C1 and C2 to the cathode of V1 180° out of phase, thus canceling. However, the desired received pulses from the converter are transferred in phase and therefore add at this point. Transformers T1 and T2 are resistively loaded to obtain a 3-db bandwidth of 10 megacycles. Capacitors C5 and C8 are factory-adjusted capacitors. Capacitors C4 and C7 provide an r-f ground (decoupling) for the plate circuits of the two stages. Resistors R5 and R6 and capacitors C10 and C11 form a decoupling (filter) network to prevent r-f signals from entering the power supply.

(2) The network composed of inductances L2 through L7 and capacitors C12 through C17, C23, and C24 provide filtering for the metering and ground return circuits for the two crystals in the balanced converter. This prevents loading of the crystal circuits and eliminates the alternating component of the intermediate frequencies. Filtering in the filament circuits of V1 and V2 prevents interaction between the two stages through the respective filaments.

(3) The power gain of this two-stage preamplifier is around 23 decibels. The output of the preamplifier is taken from i-f output jack J2 and sent through the slipring assembly GS-15536 to a 250-foot coaxial cable. This cable presents an impedance of 50 ohms to the preamplifier. The impedance of the output transformer T2 is 50 ohms to match the cable impedance. Since all adjustments within this unit are made at the factory, no field adjustments are necessary. Since the signal-to-noise ratio obtained from the grounded-grid triode depends critically upon its operating point, automatic gain control (AGC) is not used. If the AGC were employed, the signal-to-noise ratio would vary with the gain control voltage, which is undesirable. The signal-to-noise ratio obtained from this preamplifier is about 3 to 3½ decibels. The bandwidth is about 10 megacycles between the 3-db down points.

26. SUM I-F PREAMPLIFIER (fig 6-8)

Since the upper portion of the sum i-f preamplifier is identical to the error i-f preamplifier, no discussion will be made. The lower portion (AFC channel), however, differs and will be covered in detail.

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a. Block diagram discussion.

(1) V3 amplifier. The first stage of the i-f amplification is a grounded-grid amplifier similar to those discussed in paragraph 25a(1). The output is transformer-coupled to amplifier V4 and crystal delay line DL1.

(2) Crystal delay line DL1. The crystal delay line was integrated into the circuit to transform the 60-mc electrical pulses into pulses representing 2,000 yards in range. The purpose of this delay is to set the zero point of the ranging system.

(3) Amplifier V4. Amplifier V4 is a conventional amplifier identical to the operation of the second stages of i-f amplification in the error i-f preamplifier.

(4) R-F filter. Refer to paragraph 25a(3).

b. Detail circuitry discussion. The track sum i-f preamplifier GS-15691 performs the same function as the azimuth and elevation preamplifiers discussed in paragraph 25. In addition, this unit incorporates two other features. Discussion of these two features and reference to paragraph 25 above suffices for an explanation of this unit.

(1) Refer to figure 6-8. The 60-mc AFC signal taken from the balanced converter is amplified in this unit and passed on to the track AFC unit. The signal is obtained from crystals CR1 and CR2 in the AFC converter. Springs contacting crystal holders E3 and E4 connect the signal through capacitors C27 and C28. At this point converter noise cancels because of their 180° phase difference. However, the desired signals are in phase and add. These signals are connected to the cathode of tube V3. The signal is amplified by V3 and coupled across transformer T3 to delay line DL1 and tube V4. The function of DL1 is explained below in subparagraph (2). The signal is amplified by V4, coupled across transformer T4, and passed through resistor R14. It leaves the unit through jack J3 and passes to the track AFC unit. Resistor R17 is used to load the secondary of transformer T4 to provide a 10-mc band-pass transformer. The filter circuit composed of L13 through L17, C23 through C26, C31 through C33, and C36 filters the metering and ground return circuits of the two crystals (AFC) in the balanced converter. This prevents loading of the crystal circuits and eliminates (decouples) the alternating component of the i-f signal for metering purposes. The outputs for the AFC metering are applied through pins 7 and 9 of jack J1.

(2) ~~Because the transmit-receive switching is not 100 percent efficient, a small "spill over" of the main transmitter pulse surge is put into the~~

receiver. This 60-mc main surge, in addition to the 60-mc AFC signal, is amplified by V3. When this main surge reaches the delay line, the crystal transducer at the input of the delay line transforms it into a sound wave of supersonic frequency. The sound wave travels down the quartz delay line, then bounces off the highly polished end and reflects to the crystal transducer. The crystal then transfers this sound wave back into a 60-mc electrical pulse. The length of this delay line is such that the delay represents approximately 2,000 yards in range. After each firing of the magnetron, the delay line sends a series of pulses spaced 2,000 yards apart in range. A series of pulses is obtained since each reflected pulse sends another pulse down the delay line as well as an output pulse. These pulses are amplified by tube V4 and coupled across transformer T4 to the contacts of relay K1. When relay K1 is energized, these delayed pulses are connected to the output of the sum preamplifier. The purpose of this delay action is to set the zero point of the ranging system. When in the ZERO position the RANGE CALIBRATE-NORMAL-ZERO switch S3 on the target track control drawer GS-15585 applies ground to relay K1. The TEST-OPERATE switch S12, also on the target track control drawer, must be in the TEST position for this function to be accomplished. Thus the coordinated action of these two switches introduces the delay line into the sum channel and consequently the series of 2,000-yard pips to the ranging system.

27. TARGET AFC UNIT

The purpose of the target automatic frequency control unit D-153964 is to maintain a 60-megacycle frequency difference between the target radar transmitter and the local oscillator frequency. The receiving system i-f amplifiers are designed to operate at 60 megacycles, and only have a 10-megacycle bandwidth. If the difference frequency should drift more than 5 megacycles in either direction from the center frequency of 60 megacycles, signal loss in the i-f amplifier would become prohibitive. Factors which would contribute to frequency shift occur in both the transmitting system and the local oscillator, and result from a variety of causes. A drift as little as one-tenth of one percent in the frequency of either the transmitter or local oscillator frequency will change the intermediate frequency so much that the i-f amplifier stages will no longer be able to pass the signal. Under these circumstances the need of an AFC for the local oscillator becomes evident. Shifts in transmitter frequency are unavoidable, therefore the AFC, to be effective, must use a sample of the transmitted r-f energy as a reference for control of the local oscillator frequency. In the AFC mixer of the converter assembly, samples of the transmitted r-f energy are mixed with the local oscillator output and the resulting i-f signals are supplied to the AFC section of the sum preamplifier, which in turn sends 2,000-yard pulses of the i-f signal to the AFC unit. The AFC unit monitors the i-f signal, detecting any deviation from the required 60 megacycles. When a deviation

occurs, the AFC unit develops an output signal which acts to retune the local oscillator to a frequency 60 megacycles greater than the transmitted frequency. The local oscillator has a tuning triode whose elements are temperature sensitive to current flow, as determined by the output from the AFC unit. The i-f deviations, from 60 megacycles, change the current flow in the tuning triode, which in turn change the characteristic of the temperature sensitive elements. The temperature sensitive elements contract or expand changing the size of the local oscillator tuning cavity to bring the intermediate frequency back to 60 megacycles. When the target-tracking radar system is first energized the transmitter and local oscillator may have any relationship to one another. It is the function of the AFC unit under these circumstances to cause the local oscillator to search the X-band frequency range from 8,560-9,600 megacycles for the upper sideband frequency (magnetron frequency plus 60 megacycles). Once this frequency is found the AFC unit will hold the local oscillator in the constant 60-megacycle relationship as explained above. If for some reason during operation the 60-megacycle difference is lost, the AFC unit will resume its search operation until the local oscillator again locks on the correct sideband.

a. _Components._ The target radar AFC circuit may be divided functionally into seven stages: oscillator V7; multivibrator V6 as a search voltage generator, including pulse selectors V5A and V5B; i-f amplifiers V1 and V2; discriminator V3 and video amplifier V4 (excluding the diode section of V4, which must be considered with the back-bias circuit); multivibrator V6 as a tracking voltage generator; local oscillator bias control V8A; and i-f amplifier (AFC unit) back-bias control V8B.

b. _Inputs and outputs._ The input to the AFC unit is the i-f signal from the AFC channel of the sum preamplifier. The signal is applied to the grid input section of V1 on jack J1. The output to the grid of the klystron tuning triode on jack J2 is a rectangular waveform with a +15-volt to a -60-volt excursion. There are two test points, TP1 and TP2, for the purpose of checking the back-bias voltage and the operation of limiter V4.

c. _Block diagram discussion._ See figure 6-3. Operation of the target AFC breaks down into two distinct phases, the search phase and the control phase. In the search phase, V6, which actually drives the local oscillator tuning triode, is controlled by oscillator V7. In the control phase V7 is locked out and V6 is controlled by the AFC i-f signal. Oscillator V7 puts out a signal having a repetition rate of one pulse every 7.5 seconds. These pulses are applied to V6 through pulse selectors V5A and V5B. V6 is a flip-flop or bistable multivibrator which is triggered by the pulse from V7. The output from V6 is a rectangular waveform having a half-cycle period of 7.5 seconds during search operation. The output is +15 volts on the positive swing and -60 volts on the negative swing. Triggering of V6 is determined by which section is conducting at the time the trigger occurs. On the positive 15-volt swing of V6 output, the tuning triode causes the local oscillator frequency to decrease. On the negative 60-volt

swing, the tuning triode causes the local oscillator frequency to increase. One 7.5-second local oscillator search period covers the frequency range from 8,560 to 9,660 megacycles, with the next 7.5-second period covering the range from 9,660 to 8,560 megacycles. At the time the local oscillator is searching from 8,560 to 9,660 megacycles, a portion of the -60 volts is coupled back to i-f amplifiers V1 and V2 through bias control tube V8 to prevent the AFC from locking on an i-f 60 megacycles below the transmitter frequency. With V1 and V2 effectively blanked out on the negative swing of V6 output, the AFC circuit can only take over control of the local oscillator when the frequency is searching downward on the positive V6 output signal. When the AFC circuit locks on the proper frequency, oscillator V7 becomes locked out so its signal is no longer effective in the control of V6, though V7 continues to operate as before. Control of the AFC circuit is then taken over by the AFC mixer in the balanced converter. The AFC i-f signal is developed by heterodyning the local oscillator frequency and the transmitted pulse sample and is furnished to the AFC circuit through the AFC section of the sum i-f amplifier. Since V1 and V2 are biased below cutoff on the 8,560- to 9,660-megacycle sweep of the search signal, a 60-megacycle i-f signal cannot affect their operation during the upward search period. On the search sweep from 9,660 to 8,560 megacycles when the i-f signal comes within the 60 ±5 megacycle bandwidth of V1 and V2, V1 and V2 are amplified by an amount sufficient to assume control of the AFC circuit. Signals outside this 60 ±5 megacycle bandwidth are not amplified enough to couple into discriminator V3. V3 is a frequency sensitive detector that gives positive or negative video output pulses, depending on whether or not the intermediate frequency is above or below 60 megacycles. When the i-f signal is exactly 60 megacycles there will be no output from discriminator V3. If the frequency is below 60 megacycles the output will be a negative pulse. The amplifier section of V4 amplifies and inverts the discriminator video pulses, and applies them as trigger pulses to multivibrator V6. These triggers then drive multivibrator V6, causing the local oscillator frequency to go up or down depending on whether the intermediate frequency is above or below 60 megacycles. The local oscillator sweeps a very small portion of the frequency spectrum, maintaining the difference between the local oscillator and transmitted frequency well within the ±5 megacycle bandwidth of the i-f amplifiers. AFC tracking indicator I1 is a neon lamp which indicates the status of operation of the AFC circuit. During search one plate of I1 glows brightly during the 7.5-second upward search of the local oscillator and the other plate glows dimly during the 7.5-second downward search. In the control phase I1 flickers at a rapid rate. The resistive network in the output of V6 is a complex voltage divider network which establishes the +15-volt and -60-volt output signal to the local oscillator tuning triode grid and the amplitude of the signal into bias control tube V8.

    d.   Detail circuit discussion (fig 6-4).

        (1)   Oscillator V7. V7 is a JAN 6AS6 vacuum tube connected as a phantastron delay circuit. The 6AS6 is a miniature sharp-cutoff pentode whose control

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grid and suppressor grid can be used as independent control elements, and is therefore especially adaptable for use in the phantastron delay circuit. Oscillator V7 controls the search control time of multivibrator V6 and provides negative pulses at 7.5-second intervals. These pulses are differentiated by capacitor C18 and resistor R12 in the cathode circuit of V5A and by C20 and R25 in the cathode circuit of V5B. The negative spikes, corresponding to the leading edges of the pulses, pass through V5A and V5B and are the triggering signals for multivibrator V6. A phantastron delay circuit is very similar in operation to a multivibrator. The primary advantage of the phantastron is the accurate timing of the output pulses. This accurate timing is provided in oscillator V7 by regulating the discharge rate of capacitor C23 with a negative feedback from the plate to the grid. This negative feedback causes capacitor C23 to discharge at very nearly a linear rate throughout its discharge period. The actual rate at which capacitor C23 discharges is determined by the setting of search time control potentiometer R39. Operation of the phantastron can best be explained in the following manner: When voltage is first applied to the circuit, current starts to flow through V7 with a resulting drop in plate potential. (See figures 30 and 31 for circuit details and waveforms.) At the same instant there is a surge of control grid current which charges capacitor C23. The drop in plate voltage is coupled to the control grid of the tube causing the grid current to cut off, at which time capacitor C23 will start discharging through resistors R33 and R39. As capacitor C23 discharges, the voltage on the control grid tends to rise toward zero at the normal exponential discharge rate. However the rise in grid voltage results in a further increase in plate current with a corresponding drop in plate voltage. This drop in plate voltage when coupled back to capacitor C23 causes C23 to discharge at a linear rather than an exponential rate. Because of the large value of the plate load resistor R32, only a small increase of plate current is required to give the required linearity control for C23. While there is a tendency for the control grid current to continue rising, the negative feedback limits the rise to a very small amount during the initial part of the duty cycle. Because capacitor C23 discharges at a constant rate, the plate current drops linearly. This process continues until the plate current is so low that the control grid no longer controls the flow of current to the plate. At this time the screen current increases sharply, causing the screen voltage to drop due to screen current flow through resistors R30 and R31. A portion of the screen voltage drop is applied to the suppressor grid through capacitor C22, cutting the plate current off completely. When plate current no longer flows, plate voltage rises and feedback from the plate causes the control grid to rise. When the control grid becomes positive, capacitor C23 again becomes charged by control grid current flow. Capacitor C22 initially became charged by drawing suppressor grid current at the same time capacitor C23 became

charged. There was no chance for C22 to discharge during the initial phase of operation. When the screen grid pulse drives the suppressor grid negative, capacitor C22 starts to discharge through resistor R34. As C22 discharges, the voltage drop across R34 decreases until plate current can again flow. Once plate current starts to flow, the entire cycle is repeated again. The negative screen grid pulse at the end of each control grid cycle is applied to V5A and V5B for control of V6. The output, then, is a series of negative pulses, one occurring each 7.5 seconds.
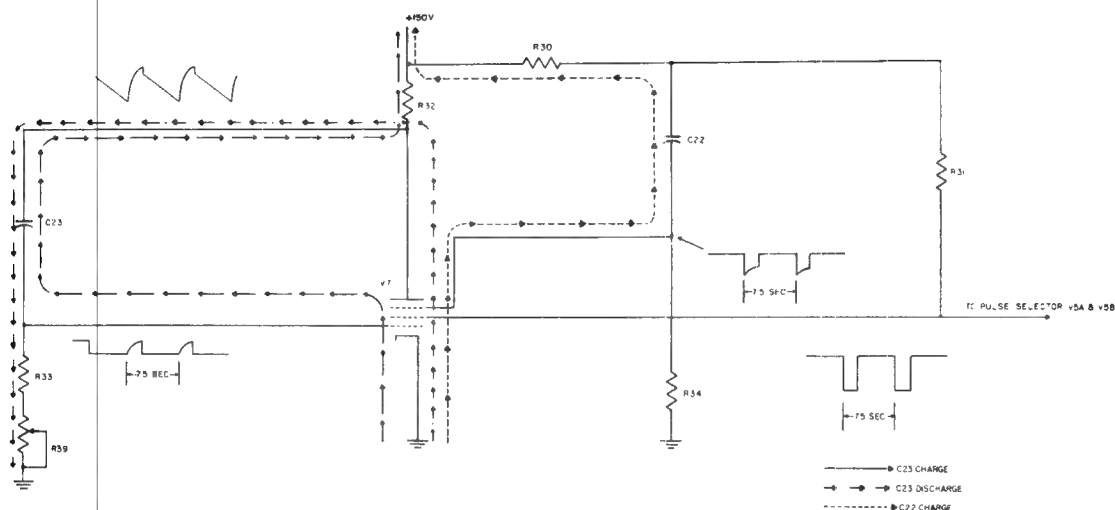
Figure 30. Phantastron oscillator, simplified schematic.

(2) Pulse selectors V5A and V5B. The pulse selector tubes V5A and V5B. are switch tubes, controlled by the conducting conditions of V6. The switching action allows the trigger signal from V7, which is applied to both V5A and V5B, to trigger the conducting half of V6, without affecting the nonconducting half. The cathode of V5A is tied back to the plate of the 6-7-8 section of V6 through the voltage divider consisting of resistors R11 and R12. The cathode of V5B is tied to the other plate of V6 through voltage divider R23 and R25. The quiescent plate voltage of V5A is established by the division of +320 volts across voltage divider R13 and R14, and is the same as grid 7 of V6. The quiescent plate voltage of V5B is developed between the plate of section 6-7-8 of V6 and the midpoint of voltage divider R18 and R19. This voltage is the same as that for grid 3 of V6. With section 6-7-8 of V6 conducting there is a relatively high-

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voltage drop across resistor R16 in the plate circuit of V6. This plate voltage drop causes the potential on the cathode of V5A to drop to a value which will permit V5A to conduct when the negative trigger is applied from V7. At the same time the cathode of V5B has the high positive voltage from the plate of nonconducting section 2-3-4 of V6 applied to it and cannot conduct when the negative trigger from V7 is applied. Conduction of V5A, when the trigger is applied, triggers the multivibrator (V6) and causes the 6-7-8 section to cut off and the 2-3-4 section to start conducting. The condition is now reversed so the B section of V5 can be used to trigger V6, but the A section cannot.

PLATE VOLTAGE

CONTROL GRID VOLTAGE

SCREEN GRID (OUTPUT) VOLTAGE

SUPPRESSOR GRID VOLTAGE

C23 DISCHARGE TIME

7.5 SEC

C22 DISCHARGE TIME
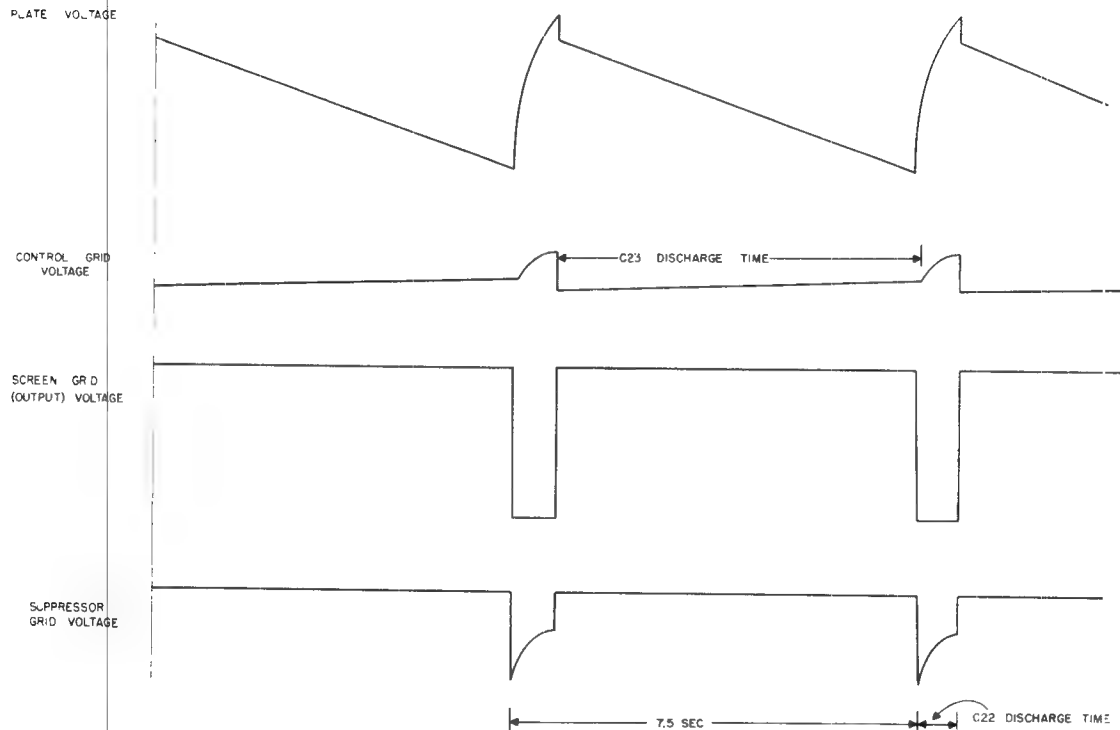
Figure 31. Phantastron waveforms.

(3) Multivibrator V6 (as a search voltage generator). V6 is a bistable multivibrator, or flip-flop circuit. With one section, conducting the voltage across cathode resistor R15 is sufficient to hold the opposite section cut off. Coupling between plate 6 of V6 and grid 3 of V6 is through the parallel combination of capacitor C19 and resistor R18. Cathode coupling is used for cutoff of the 6-7-8 section of V6. Grid voltage for each section is established by voltage dividers. Resistors R13 and R14 establish the grid-to-cathode voltage of grid 7. Resistors R17, R18 and R19 establish the grid-to-cathode voltage of grid 3. Resistor R16 is the plate load for section 6-7-8 of V6 and resistor R17

is the plate load for the opposite section. Resistors R20 and R21 are in series with the tracking indicator I1, across the plates of V6, and are used as current limiting elements. When section 6-7-8 of V6 is conducting, section 2-3-4 is held at cutoff by the voltage drop across resistor R15 in the cathode circuit. A trigger from V7 will drop the potential on grid 7 of V6 which will in turn raise the plate 6 potential. The rise in plate voltage is applied to grid 3 causing section 2-3-4 of V6 to commence conducting. The conduction through section 2-3-4 increases the voltage drop across cathode resistor R15 further increasing the voltage difference between the cathode and grid 7. This regenerative feedback cycle continues until section 6-7-8 is completely cut off and section 2-3-4 is conducting heavily. This condition of operation will continue until the next trigger pulse from V7 occurs 7.5 seconds later. The next trigger causes grid 3 to go negative, reducing current flow through section 2-3-4 of V6. The resulting voltage drop across cathode resistor R15 causes section 6-7-8 to start conducting. A drop in plate 6 voltage results, which, when coupled to grid 3, causes section 2-3-4 to cut off and section 6-7-8 to conduct at its maximum rate. The output, a 4 cycle per minute rectangular waveform, is coupled to the grid of the tuning triode in the local oscillator through a complex resistive network that makes the reference voltage give a signal swing of +15 to -60 volts to the reference oscillator. The swing in the positive direction is limited to +15 volts by the limiting action of section 6-7-8 of V8 and can be adjusted by bias control potentiometer R36.

(4) I-F amplifiers V1 and V2. I-F amplifiers V1 and V2 provide the amplication necessary to raise the amplitude of the i-f signal to a usable value for AFC tracking. Each stage is broad tuned to give a 10-mc bandwidth (60±5 megacycles). Any signal of the required amplitude within the 60±5-mc frequency range will be amplified by V1 and V2. Signals outside that frequency range will not be amplified and will be of insufficient amplitude for use in AFC tracking. During the search phase the amplifiers have no way of determining if the local oscillator frequency giving the i-f difference frequency is 60 megacycles above or below the transmitter frequency. Back-bias is therefore fed back from the bias control circuit to cut V1 and V2 off during the upward search of the local oscillator. V1 and V2 start conducting when the downward search begins and will only lock on an i-f signal developed when the local oscillator is 60 megacycles above the transmitter frequency. The input to V1 comes into jack J1 through a coaxial cable. Resistive networks R48, R49, and R50 match the primary impedance of transformer T1 to the impedance of the delay line and serve to broaden the frequency response of the series combination capacitor C1 and the primary of transformer T1. Resistors R49 and R50 also form a voltage divider which reduces slightly the amplitude into capacitor C1 and transformer T1. The i-f signal in the primary of transformer T1 is transformer coupled to the grid of V1. The secondary

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of transformer T1 is heavily loaded by parallel resistor R1 to increase
its bandwidth to the required 10 megacycles. D-C plate and screen
voltage is established by the voltage drop across resistor R37, which
is decoupled by capacitor C4. The d-c voltage drop across the primary
of transformer T2 is negligible so the plate and screen voltage are
essentially the same value. The primary of transformer T2 is paral-
leled by bandwidth band-broadening resistor R3. Cathode bias is pro-
vided by plate current through resistor R2 and capacitor C2 in the cathode
circuit of V1. V2 operation is identical to that of V1. The only dif-
ference lies in the lack of bandwidth band-broadening elements across
the transformer windings of T2. These elements are not required be-
cause of the high amplification through V1. Resistor R38 is the plate
and screen dropping resistor for V2. Capacitor C8 provides the neces-
sary decoupling for the plate and screen voltage. Cathode bias is pro-
vided by resistor R5 and capacitor C6. In addition to cathode bias, the
back-bias control is coupled to the grids of both V1 and V2 from the
bias control circuit. The i-f is blocked out of the back-bias circuit by
the filtering action of inductance L9 and capacitor C6 in the grid circuit
of V1 and inductance L3 and capacitor C5 in the grid circuit of V2. Dual
coupling of the i-f signal to the discriminator V3 is provided by trans-
former coupling from the primary to the secondary of transformer T3
and by impedance coupling from the plate side of transformer T3 primary
winding to the midpoint of capacitors C9 and C10 in the discriminator
circuit.

(5) Discriminator V3. The discriminator transforms frequency changes in
the i-f to pulse amplitude voltage variation. These variations will be
either positive or negative pulses depending on whether the i-f is higher
or lower than 60 megacycles. The amplitude of the pulse is determined
by the amount of frequency deviation from the 60-mc center frequency.
The signal voltages applied to the plates of V3 from V2 consist of two
components: a reference voltage applied directly to the center of the
secondary of transformer T3 (at the midpoint of C9 and C10); and the
voltage induced in the secondary of transformer T3 by inductive coupling.
Both the primary and the secondary of transformer T3 resonate at 60
megacycles, with the two windings being loosely coupled. The voltage
Er (fig 32) represents the reference i-f voltage developed across the
primary of transformer T3 and applied through capacitors C9 and C10
to the secondary of transformer T3. The Ei represents the total
induced i-f voltage in the secondary of transformer T3. At the resonant
frequency of 60 megacycles, the induced voltage is 90° out of phase with
the reference voltage. Half of Ei appears across capacitor C9 and the
other half appears across capacitor C10. Resultant voltage E1 is
applied to the upper diode of tube V3 and resultant voltage E2 is applied
to the lower diode of V3. In the figure, voltages Er, E1, Ei, and E2
are represented by vectors. At 60 megacycles the resultant voltages

E1 and E2 are equal in amplitude. Therefore both sides of V3 conduct equally, but their current flows are equal and opposite with respect to ground reference (currents through resistors R6 and R7 are equal and opposite). When the i-f signal is below 60 megacycles, the capacitive reactance in the secondary of transformer T3 is greater than the inductive reactance and the circuit is no longer at resonance. The induced voltage Ei is no longer 90° out of phase with the reference voltage Er, and as indicated by the vector addition of Ei and Er voltage, E1 is now greater than voltage E2. This produces greater conduction in the upper diode of V3, more current flow in resistor R7 than in R6 (polarities indicated on figure 32), and a positive video pulse as the output of the discriminator. When the i-f is above 60 megacycles the situation reverses, resulting in more conduction through resistor R6 than through R7. The output is then a negative video pulse. The i-f is changed to video pulses by a rectifying and filtering action. The diodes can only conduct on the positive half of each cycle of i-f signal. Conduction of the upper half of V3 charges capacitor C14 to the peak value of the positive half cycle of the i-f signal. Capacitor C14 then discharges through resistor R6, maintaining the voltage near the peak value until the next positive swing. This filtering action maintains a nearly constant voltage across resistor R6 during application of the i-f signal. The same explanation applies to the lower half of V3 and capacitor C15. Inductances L6 and L7 are r-f chokes which react with the interelectrode capacitances of V3 to form a shunt peaking circuit, thereby holding up the effective amplitude of the i-f signal on the plates of V3. The output from the upper end of resistors R6 and R7 is applied through coupling capacitor C16 to the grid of the video amplifier section of V4. The filament circuits of V1, V2, and V3 have r-f chokes in series with them and are bypassed by 5,000-micromicrofarad capacitors to prevent i-f feedback into other stages of the system.

(6) Video amplifier and diode V4. V4 consists of a video amplifier section and a diode section. The video amplifier amplifies and inverts the video pulses from the discriminator to furnish pulses of the proper amplitude and polarity to control multivibrator V6 during local oscillator frequency tracking. Capacitor C16 and resistor R8 are a conventional coupling circuit for the pulses from the discriminator. Resistor R47 is a parasitic suppressor which prevents oscillations in the grid circuit due to lead inductance resonating with grid-to-cathode capacitance. These oscillations which tend to occur on the leading and trailing edge of the video pulses might trigger V6 prematurely. Resistor R10 is the plate load resistor from which the video output is coupled through capacitor C17 to grid 7 of V6. Resistor R9 is the bias resistor for V4 and has a no-signal voltage developed across it of +3 volts. The diode section of V4 is used to limit the positive swing of the back-bias control voltage

to the +3 volts on the cathode. Any time the back-bias voltage exceeds +3 volts, the 2-3-4 section of V4 conducts, heavily limiting the rise to the existing cathode voltage.
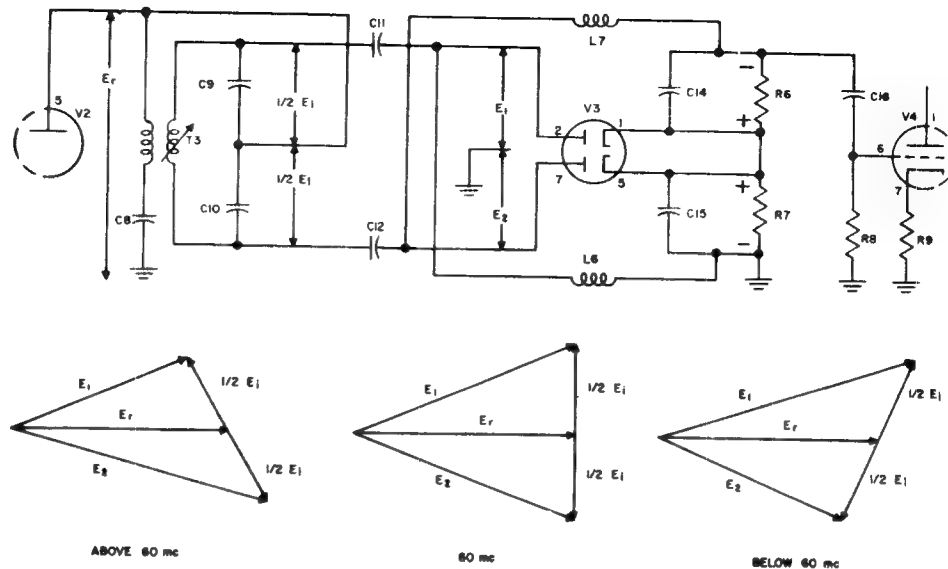
Figure 32. Discriminator circuit, simplified schematic.

(7) Multivibrator V6 (as a track voltage generator). Once the AFC circuit has locked on the proper intermediate frequency, oscillator V7 is no longer effective in control of V6. The control of V6 is taken over by the video pulse coming in from discriminator V3. These pulses occur at a rate of 1,000 pulses per second. The polarity of the pulses is determined by the relation of the intermediate freqeuncy to the 60-mc center frequency of the discriminator circuit. When the intermediate frequency is exactly 60 megacycles there will be no output pulses from the discriminator. V6 control of the tuning triode grid causes either an upward or downward search of the local oscillator frequency depending on which half of V6 is conducting. The cooling rate of the control element in the tuning triode is relatively low with respect to the 1,000 pulse per second input to V6. As a result, during track, V6 control of the local oscillator keeps the local oscillator frequency within very close tolerence and maintains the i-f signal at, or very near, 60 megacycles. To illustrate how V6 controls local oscillator tracking, assume the following sequence of operation: section 6-7-8 of V6 is cut off, section 2-3-4 is conducting; applying a negative 60 volts to the tuning

triode grid. This negative voltage applied to the tuning triode will cause the local oscillator to search upward in frequency. Also assume that at this time, the i-f signal is slightly below 60 megacycles. The input pulse to grid 7 of V6 will then be a negative pulse. Since section 6-7-8 of V6 is already cut off, the negative pulse will have no effect on V6 operation and the local oscillator will continue to search upward in frequency. At the end of the next 1,000 microseconds suppose the i-f is exactly 60 megacycles. The output from the discriminator will be zero, so V6 will again keep searching upward. After another 1,000 microseconds the local oscillator will have gone very slightly above that required to give a 60-mc intermediate frequency, so the signal to grid 7 of V6 will be a positive pulse. This positive pulse will cause section 6-7-8 of V6 to conduct and will cut off section 2-3-4, causing the local oscillator frequency to start downward. Since the heating and cooling rate of the temperature sensitive element in the tuning triode is very slow, switching of V6 does not occur on every pulse. This may result in 3 or 4 pulses occurring before the local oscillator frequency becomes low enough to cause V6 to switch and give another upward swing of the local oscillator frequency. The end result is a very small frequency deviation well within the ±5-mc bandwidth of the i-f amplifier. V7 may momentarily affect the operation of V6. Occasionally, however, the i-f signal immediately assumes control again so it has no effect on the over-all track operation.

(8) Bias control V8. The bias control section of V8 is comprised of the 2-3-4 section. The purpose of the bias control is to cut off the i-f amplifiers during the upward swing of the local oscillator frequency and allow V1 and V2 to conduct on the downward swing. Then when the AFC system locks on the proper frequency the bias control must not interfere with AFC tracking. Operation of section 6-7-8 of V8 in establishing the +15-volt limit of the AFC output signal has already been explained. Back-bias control is a function of the conducting properties of section 2-3-4 of V8, the time constant of capacitor C24 and resistors R40 and R43. During the search phase the downward local oscillator frequency sweep period is 7.5 seconds and the upward sweep is 7.5 seconds. The -60-volt signal that controls the upward sweep is applied to the grid of V8 through capacitor C24 and resistor R40. This voltage causes grid 3 of V8 to become negative with a resulting rise of plate voltage making the back-bias voltage positive. This condition is only momentary, however, due to the length of the 7.5-second period as compared to the time constant of capacitor C24 and resistors R40 and R43. The back-bias returns to the negative potential established by voltage divider R44, R45 and R41 giving a negative voltage of sufficient amplitude to cut off V1 and V2. When the output voltage goes positive to start the downward search, V1 and V2 are unblocked by a positive voltage through resistor R26 which overcomes the blocking bias. The

amplitude of the positive voltage is limited by diode section 2-3-4 of V4.
During frequency tracking of the AFC circuit the repetition rate of the
multivibrator output signal is short compared to the time constant of
capacitor C24 and resistors R40 and R43. This combination then
becomes a coupling circuit and V8 prevents the negative signal swing
from cutting off V1 and V2. The negative swing coupled to grid 3 of V8
causes its plate to go positive, raising the voltage at the junction of
resistors R44 and R45 and keeping V1 and V2 in a conducting state.

28. TARGET AGC UNIT (fig 6-18)

The output from the sum i-f main amplifier is used as a reference voltage,
against which the phase and amplitude of the azimuth and elevation error signals
are compared. By signal comparison, the system is capable of determining both
the amplitude and direction of antenna pointing error. To provide reliable error
information, the comparison (sum channel) voltage must be constant. The AGC
channel provides bias control voltages which vary the gain of the sum i-f main
amplifier to provide the necessary constant i-f main amplifier signal amplitude.
AGC also furnishes proportional bias voltage to the azimuth and elevation i-f
main amplifiers to control the gain in such a manner as to provide a constant
volt-per-mil output regardless of change in signal strength. However, the AGC
bias level to the azimuth and elevation i-f main amplifiers does not maintain a
constant signal amplitude for these two channels. The amplitude of the azimuth
and elevation signals vary in amplitude by amounts proportional to the antenna
pointing error. The AGC unit is not an instantaneous gain control. Several
received video pulses are required to effect a change in the gain of the i-f main
amplifiers. The AGC unit has two inputs; one is a 0.4μsec, 15-volt gate
pulse from the ranging system, the other is the 0.3μsec AGC video pulse
from the video and phase unit. The AGC video pulse amplitude varies with the
strength of the received signal. The output from the AGC unit is a bias control
voltage proportional to the received signal strength used to vary the gain of
stages 2 through 6 in i-f main amplifiers GS-15589.

a. Block diagram discussion. See figure 6-17. The AGC unit output (control
bias) voltage amplitude is a function of the amplitude of the received target echo,
therefore the input to the AGC unit is target video pulses from the sum channel
of the video and phase unit. This information is brought into cathode follower
V3A on AGC VIDEO jack J2. The video signals into V3A are all the video signals
within the range of the radar at the target azimuth. To limit AGC action to the
required target video pulse, the receiver gate pulse is used to gate the AGC unit.
The receiver gate is varied by the range unit to gate any desired video signal
within the range of the radar and exclude all others. The gate pulse is applied to
gate amplifier V1 through RECEIVER GATE jack J1. From V1 the signal goes
to cathode follower V2 and then to signal gating tube V4 and gated detector V6.
The actual gating, or selection, of the target is accomplished in gating section
V4 by slewing the radar in range until the receiver gate is coincident with the

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desired target video signal. The gated video signal will appear in the output of V4 as a negative pulse with one half the amplitude of the input pulse. The negative signal is then amplified and inverted by V5A and integrated in the plate circuit to stretch the positive output pulse from 0.3 to 10 microseconds wide. The stretched positive pulse is next amplified by video amplifier V5B and applied to gated detector V6. The output from V5B to V6 is a negative 10μsec pulse. The gated detector V6 uses the receiver gate to cut the width of the stretched video pulse back to 0.4 microsecond and to provide a d-c output voltage whose amplitude is proportional to the amplitude of the target video pulses. From V6 the d-c voltage is applied to comparator V7, where it is compared to a negative voltage from the threshold potentiometer. The threshold potentiometer voltage may be varied to change the amount of V6 output voltage required to give AGC action. Provisions are also made to allow manual gain control of the i-f amplifiers. The output from V7 is filtered to prevent any tendency toward oscillation and still permit relatively fast AGC action. D-C amplifier V8 amplifies the output from comparator V7. The SUM ZERO control sets the bias on V8 to give a zero output to cathode followers V9, V10, and V11, with a zero input video signal to the AGC unit. Crystal CR1 in the output circuit of V8 is used to compensate for nonlinearity of the gain versus grid-voltage relationship of the i-f main amplifiers. Crystal CR2 limits the positive voltage at TP2 to zero volts. The AGC voltage to cathode followers V9, V10, and V11 is a negative d-c voltage proportional to the amplitude of the input video signal. Cathode follower V9 has a SUM LEVEL control which allows the AGC bias level to be adjusted to maintain the amplitude of the sum i-f main amplifier signal constant. V9 and V10 have slope controls to establish a gain level for the azimuth and elevation channels that is 3 db below the level of the sum channel and on a curve parallel to the sum curve. The azimuth and elevation channels are AGC controlled, but not to the extent that they will maintain the azimuth and elevation i-f main amplifier output pulses at a constant amplitude. AGC metering is provided through cathode follower V3B to test meter M1 in the target test panel. Adequate switching is provided to allow monitoring of all three AGC outputs. V3B provides AGC information to the event recorder and to the AGC monitor when this unit is used in the missile-tracking radar.

b. Detailed circuit operation (fig 6-18).

(1) Gate amplifier V1. The gate amplifier consists of sections A and B of V1. The receiver gate pulse at J1 is coupled to the grid of V1A through capacitor C1. The grid bias for V1A is obtained by tapping off at the midpoint of voltage divider resistors R2 and R5. Resistor R1 establishes the plate voltage for V1A. Plate voltage is kept constant by the filtering action of capacitor C2. The positive 0.4μsec pulse on the grid of V1A causes an increase in the voltage drop across common cathode resistor R3. The rise in voltage across resistor R3 results in a decrease in current through V1B, causing the plate voltage of V1B to go positive. Cathode coupling of this nature allows amplification with minimum distortion. Resistor R6, the plate load resistor of V1B, is returned to

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resistor R1 for its B+ voltage. Grid voltage for V1B is provided by resistor R4 in the cathode circuit. The grid voltage is held constant by bypass capacitor C4. The output from V1B is a positive pulse of approximately 30 volts amplitude coupled to both grids of cathode follower V2 through capacitor C3.

(2) Cathode follower V2. Both sections (A and B) of V2 are connected as a cathode follower giving two outputs, one across resistor R10 in the A section and one across resistor R11 in the B section. The 30-volt positive pulse from V1B is developed across grid load resistor R8 and applied to both grids of V2. Resistor R7 in series with the grid of V2A and resistor R9 in series with the grid of V2B suppress parasitic oscillations in the cathode follower circuit. Plate voltage for V2 is established by dropping resistor R100, and the plate voltage is held constant by decoupling capacitor C21. Cathode followers are used to match the output impedance to the primary impedance of transformers T1 and T2. The output pulse developed across resistor R10 is coupled through capacitor C5 to the primary of transformer T1 in the signal gating stage V4. The output developed across resistor R11 is coupled through capacitor C6 to the primary of transformer T2 in the V6 detector stage.

(3) AGC video amplifier V3A. Tube V3A is a cathode follower used as a buffer to isolate the gating action of tube V4 and transformer T1 from the sum i-f channel in the video and phase unit. The AGC video signal is developed across resistor R96 and applied to the grid of V3A through resistor R97. Resistor R97 is a parasitic suppressor. Bias for V3A is developed across resistor R99 from the -250 volts supplied across voltage divider resistors R98 and R99. The gain of V3A is almost unity so that the output signal coupled to resistor R14 through capacitor C22 is almost the same amplitude as the input pulse. The average pulse amplitude is approximately 1 volt, however the amplitude varies with received signal strength. The pulse developed across resistor R14 is applied to terminals 2 and 3 of the secondary of transformer T1 in the V4 signal gating circuit.

(4) Signal gating circuit V4. V4 is a switching circuit used to provide the gating action for the AGC video pulse. Both sections of tube V4 are held in the cutoff condition by the voltages across the voltage dividers in the plate and cathode circuits. V4B is cut off by the +250 volts applied to its cathode through SIGNAL GATE BAL potentiometer R15, resistors R16 and R12, and common resistor R14 at the center tap of transformer T1 secondary. V4A is cut off by -250 volts applied to its plate through resistors R17 and R13 and common resistor R14. The AGC video pulse does not have sufficient amplitude to cause V4A or V4B to conduct in the absence of a gate signal from cathode follower V2. The switching action

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of V4 is provided by the receiver gate pulse from cathode follower V2. The positive receiver gate pulse on the primary of transformer T1 is coupled to the secondary as a negative signal on terminal 4 and a positive signal on terminal 1 of transformer T1. The midpoint of transformer T1 (terminals 2 and 3) remains at ground potential. The negative pulse at terminal 4 of transformer T1 and the positive pulse at terminal 1 of transformer T1 cause equal conduction of V4A and V4B. With equal conduction of V4A and V4B the voltage output to grid 3 of V5A remains at ground potential, or at the same potential as terminals 2 and 3 of transformer T1. As a result, the receiver gate signal by itself does not affect the output to V5A, it only switches V4 into conduction so the AGC video pulses can give the desired output. The receiver gate signal, as previously mentioned, may be moved in and out in range to gate any video pulse within range of the target-tracking radar. Once the two sections of V4 are switched into conduction by the gate signal, V4A and V4B may be considered as two sections of a voltage divider, with the internal resistance of each section controlled by the AGC video pulse. The negative AGC video applied at terminals' 2 and 3 of transformer T1 causes an increase of the effective receiver gate signal across V4B, and an equally effective decrease of the receiver gate across V4A. This results in a negative signal across resistor R18 at the grid of V5A. Since the AGC signal voltage is developed across both V4A and V4B, the output pulse amplitude is one-half the input pulse amplitude. Resistor R43 serves the dual purpose of suppressing any tendency of the secondary of transformer T1 from oscillating with the distributed wiring and tube interelectrode capacitance, and of broadening the response characteristic of the transformer in order to preserve the shape of the receiver gate and AGC video pulses. Capacitors C7 and C8 are provided to permit efficient coupling of the signal gate AGC video signal to cathode 5 and plate 7 of V4. Capacitors C9 and C10 are used to compensate for stray wiring and interelectrode capacitance within the circuit. SIG GATE BAL 2 capacitor C10 is adjustable to permit balance of the capacitance of the two sections of V4. SIG GATE BAL 1 potentiometer R15 allows blanacing of the resistive elements of both halves of V4. Proper balance of the signal gating circuit is achieved when zero voltage is obtained with the receiver gate present, but with no AGC video signal. Both capacitor C10 and resistor R15 must be adjusted to obtain proper balance. With a receiver gate signal and an AGC video signal input into V4, the output to grid 3 of V5 is a negative pulse whose amplitude is proportional to the amplitude of the AGC video input pulse.

(5) Pulse amplifier and integrator V5. V5A and V5B are typical R-C coupled amplifiers, used to amplify the gated AGC video pulse to the required amplitude. Between the two amplifier stages is an integrating circuit

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which stretches, or increases, the width of the pulse. Pulse stretching is done at this stage to increase the energy content of the pulse and reduce the effect of any hash which might exist on the AGC video signal. The negative video signal on grid 3 of V5A reduces the plate current through V5A causing the plate voltage to go positive as a result of the reduced voltage drop across plate load resistors R19 and R20. When the plate voltage rises, capacitor C12 charges through resistor R21 toward the existing plate voltage. Because of the short pulse period (0.3 microsecond), capacitor C12 can only charge a short time. At the end of the AGC video pulse, plate 4 again tends to return to the lower plate voltage which existed before the signal was applied. However, the charge on capacitor C12 now becomes the effective plate voltage and prevents an immediate return to the lower value. Capacitor C12, in discharging, has an electron path through V5A and resistor R21 having a much longer time constant than its charge path through resistors R21, R19, and R20. As a result, the return to the no-signal plate voltage takes considerable time, thus stretching the pulse appreciable. The stretched pulse is coupled to the grid of V5B through capacitor C11 where it is developed across resistor R22. Resistor R48 is a parasitic suppressor. TPI is a test point provided to monitor the signal while setting the balance controls in the signal gating stage. V5B amplifies and inverts the signal, giving a negative 10μsec pulse to detector stage V6. Resistor R23 is the plate load for V5B. Cathode bias for V5B is provided by resistor R24. R24 is unbypassed to insure more faithful reproduction of the input signal. The signal is coupled to the midpoint of transformer T2 through capacitor C13.

(6) Detector V6. In many respects the operation of V6 is identical to that of signal gating circuit V4. The purpose of detector V6 is to convert the stretched AGC video signals to 0.4 microsecond, then provide a d-c output signal proportional to the peak amplitude of the incoming video pulses. In the absence of an AGC video circuit, a receiver gate signal from cathode follower V2 will couple a positive voltage to terminal 4 of transformer T2 and an equal negative signal to terminal 1 of T2. Terminals 2 and 3 of T2 will be at ground potential. The voltage across the secondary of transformer T2 will cause current flow through V6A and V6B, charging capacitor C14 negative at the plate of V6A and capacitor C15 positive at the cathode of V6B. At the end of the 0.4μsec gate pulse period, capacitor C14 will start discharging slowly through resistor R26 creating a negative voltage at plate 7 of V6A. At the same time capacitor C15 will discharge through resistor R27 developing a positive voltage at cathode 5 of V6B. These voltages hold both sections of V6 cut off until the next receiver gate signal, regardless of what is applied to terminals 2 and 3 of transformer T2 from the AGC video channel. Capacitors C14 and C15 also permit efficient coupling of the AGC video pulses to V6A and V6B. With no input from the AGC video channel the

signal gate pulse causes equal conduction of both halves of V6. The output to the grid of V7A, across capacitor C16, remains zero. A negative input pulse from the AGC video channel at the same time as the receiver gate signal will result in a greater voltage across V6A than across V6B. This causes the voltage applied to capacitor C16 and grid 3 of V7A to drop below ground potential by an amount equal to the voltage developed across resistor R25. At the end of the gate pulse, V6A and V6B immediately cut off, leaving capacitor C16 charged to the amplitude of the peak value of the AGC video pulse. C16 will maintain its charge until the next gate and video pulse. If the next gated pulse is smaller amplitude, capacitor C16 will discharge to its peak value. If the pulse is larger, C16 will charge to the peak of the larger signal. The charge across capacitor C16 then is a d-c voltage whose amplitude is dependent on the amplitude of the AGC video pulses. R101 serves the same purpose as resistor R43 in the signal gating circuit. The pulse stretching in the integrator circuit plays a very important function in the detector circuit. Suppose the integration has not been included previously and a square $0.4\mu sec$ pulse was used as the error control voltage. If, by unforseen conditions, this control voltage and the gating pulse (0.4 microsecond each) did not appear at identical times on transformer T2, the charge on capacitor C16 would be disrupted. If the gating pulse happened to last slightly longer than the control pulse, the two diodes would continue to conduct during the remainder of the gating pulse period. This would appear as a zero error to capacitor C16, causing it to discharge. Since the integration circuit is included, the control voltage is much longer in duration than the gating voltage, and the undesirable effect does not occur.

(7) Comparator V7. Tube V7 is used as a comparison circuit. The detected AGC voltage across capacitor C16 is compared with the voltage from adjustable THRESHOLD potentiometer R45 to determine the effective level of AGC voltage applied to grid 3 of d-c amplifier V8. The threshold adjustment determines the level at which the error voltage across capacitor C16 starts its effect on the i-f main amplifiers. Tube V7A and V7B are parallel-connected triodes with coupling from V7A to V7B provided across cathode resistors R29 and R30. Resistors R28 and R34 are the plate load resistors for the two sections of V7. Resistors R31, R32, and R33 are provided as a balance network to stabilize the operation of V7A and V7B. The negative AGC error voltage across capacitor C16 causes a decrease in conduction through V7A. This results in a drop in the voltage across resistors R29 and R30 causing an increase in conduction through V7B. The amount of error voltage on capacitor C16 required to affect conduction through V7B can be adjusted with THRESHOLD potentiometer R45. This potentiometer is part of a voltage divider across the -250-volt power supply. As the threshold voltage applied to grid 7 of V7A is made more negative, the required AGC error level to affect i-f amplifier gain becomes greater. The threshold voltage is

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sometimes referred to as a delay voltage. This does not mean a delay in time, but a delay in the sense that the error voltage has to reach a certain negative value before it has any effect on the gain of the i-f amplifiers, as previously mentioned. The output voltage from V7A is a negative d-c voltage which varies with respect to a fixed reference. This reference is established by voltage divider resistors R34, R35, R36, R37, and R38 between +250 volts and -250 volts. The amount by which the voltage between resistors R36 and R37 goes negative with respect to its fixed reference for a given AGC error voltage is determined by how much the error signal goes negative with respect to the threshold voltage. The AGC action is relatively fast, as compared to other radar AGC circuits, and tends to be unstable. Therefore the filter circuit capacitors C17, C18, and C19 and resistor R39 is provided to eliminate any tendency toward self-oscillation.

(8) D-C amplifier V8. D-C amplification without signal inversion of the AGC error voltage is provided by using cathode coupling from the A section of V8 to the B section. The negative d-c AGC video voltage on the grid of V8B reduces the voltage drop across R40. This reduction of cathode voltage increases plate current through V8B causing the plate voltage to go negative. The negative-going plate voltage is then applied to the three parallel AGC cathode followers V9, V10, and V11. The AGC error voltage is also applied to SUM ZERO BAL switch S2 for metering the d-c amplifier output. SUM ZERO potentiometer R58 is part of a voltage divider network which is used to adjust the fixed bias on grid 7 of V8B. The potentiometer is adjusted for a zero output to the cathode followers, with a zero input AGC video signal to V3A. Noise, hash, and different characteristics of tubes in preceding stages might vary the zero level of section V8B. Therefore the bias on the tube is made variable with potentiometer R58. When adjusting R58, a proper d-c voltage reading is obtained through use of the AGC metering circuit. Crystal CR1 is used to compensate for nonlinear-gain versus grid-voltage characteristics of the main i-f amplifiers. Crystal CR1 is biased on the knee of its characteristic curve (fig 33) to give an AGC error voltage which increases in proportion to the decrease in gain of the i-f amplifier. This tends to make the gain-grid voltage curve a straight line giving a linear output. In figure 33, curve A represents the change in gain the voltage from crystal CR1 would produce in a perfectly linear amplifier. Curve C represents the actual gain characteristic of the i-f amplifiers. Curve B shows the desired gain versus bias characteristic obtained by combining curve A and curve C. The final result obtained by use of crystal CR1 is an output from the AGC unit which will make the i-f main amplifiers linear over their control range. Crystal CR2 is a positive limiter which prevents the voltage at TP2 from going positive with respect to ground.
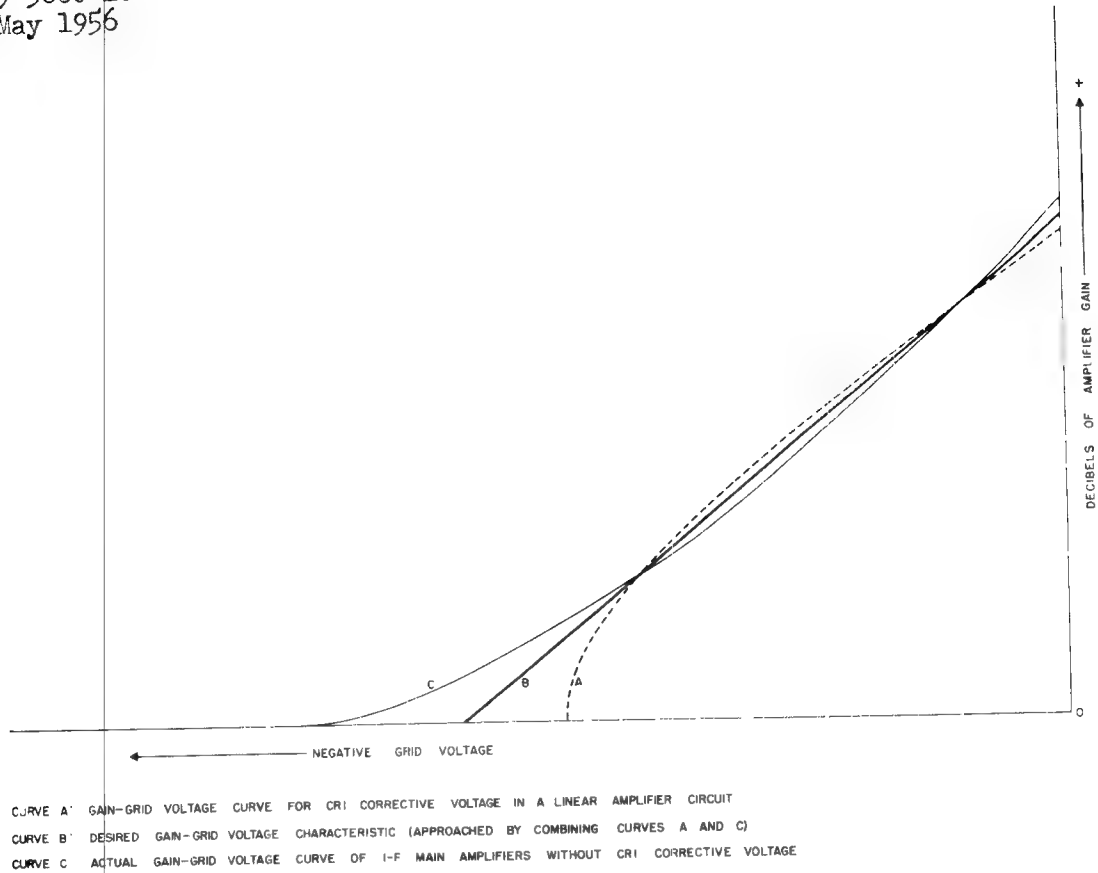
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CURVE A   GAIN—GRID VOLTAGE CURVE FOR CRI CORRECTIVE VOLTAGE IN A LINEAR AMPLIFIER CIRCUIT

CURVE B   DESIRED GAIN—GRID VOLTAGE CHARACTERISTIC (APPROACHED BY COMBINING CURVES A AND C)

CURVE C   ACTUAL GAIN—GRID VOLTAGE CURVE OF I-F MAIN AMPLIFIERS WITHOUT CRI CORRECTIVE VOLTAGE

Figure 33. Correcting action of crystal CR1.

(9)   Cathode followers V9, V10, and V11. The output cathode followers are
used to match the output impedance of the AGC unit to the grid circuits
of the i-f amplifiers. The output of these cathode followers may be
varied between the limits of 0 and minus 6 volts normal, depending on
the various potentiometer settings and the amplitude of the AGC control
voltage. Each cathode follower uses a dual triode connected in parallel.
The AGC unit input is taken from the sum AGC video channel and controls
the gain of all three i-f main amplifier channels. Since all three channels
will vary slightly in their gain, adjustments must be made to provide the
same gain characteristics for each of the three channels. SUM LEVEL,
AZ LEVEL and EL LEVEL potentiometers R68, R77, and R86 respec-
tively are used to control the fixed bias on each of the cathode followers.
This sets the gain level of the i-f main amplifiers. The gain level of the
sum i-f main amplifier is set to maintain a constant i-f amplitude. The
azimuth and elevation i-f main amplifiers are set to give gain control
throughout the effective range of the tracking radar, but they do not
maintain a constant i-f amplitude. AZ SLOPE potentiometer R74 and EL
SLOPE potentiometer R83 control the amount of AGC voltage that is applied
to the grids of V10 and V11. The slope adjustments are used to adjust the
AGC voltage along the characteristic curves of V10 and V11. The azimuth

and elevation slope controls are adjusted to cause V10 and V11 to operate along the same portion of their characteristic curve as sum cathode follower V9. The complex voltage divider inputs to V9, V10, and V11 establish operating potentials for the control elements of the cathode follower tubes. The AGC bias is applied to the grid circuits of the respective i-f main amplifiers through terminals 4, 10, and 3 of plug P2. The AGC voltages are also coupled to the AGC metering circuit through isolating resistors R73, R82, and R91.

(10) AGC metering circuit. The AGC metering circuit is used to monitor the outputs of the AGC cathode followers and the output of d-c amplifier V8. The readings obtained on the RCVR TEST meter are used in making adjustments of the AGC output voltages. The RCVR TEST meter is located in track test panel GS-15597, with the bulk of the circuitry in the AGC unit (fig 34). Tube V3B is a cathode follower circuit used to isolate the AGC circuit from the meter circuit and the event recorder which is provided with an output. The AGC outputs to be metered are taken from the cathodes of V9, V10, and V11 at pins 6, 8, and 5, respectively, of plug P2. They enter the track test panel at pins 11, 7, and 9 of plug P1. The three inputs terminate at various contacts of BIAS switch S2. Switch S2 may be positioned to allow metering of any one of the three bias voltages. The AGC voltage controls the conduction through tube V3B. The resulting proportional voltage drop across METER BAL potentiometer R93 is reflected on RCVR METER M1. SUM ZERO BAL switch S2 removes the AGC output voltage and applies the plate voltage from V8 giving a meter indication used to adjust the SUM ZERO potentiometer R58. METER BAL switch S1 grounds the grid of V3B so METER BAL potentiometer R93 can be balanced, or calibrated, to give correct indications.

(11) Manual AGC gain control. Manual AGC gain control facilitates performance of various tests which must be made on the target-tracking radar. When the MAN GAIN CONTROL relay K1 in the AGC unit is energized, the gain of the three i-f main amplifiers is controlled manually from one of two locations; the track test panel or the target track control panel. MAN GAIN CONTROL relay K1, however, cannot be energized from either control position unless the TEST-OPERATE switch S12 in the target track control drawer is in the TEST position. When MAN GAIN CONTROL relay K1 in the AGC unit is energized (fig 35), a negative voltage (less than -1 volt) is applied from voltage divider resistors R41 and R42 to terminals 2 and 3 of transformer T2. This effectively blocks out the AGC video signal and gives a negative d-c error signal across capacitor C16 at the grid of V7A which is proportional to the voltage across resistor R41. At the same time, contacts of energized MAN GAIN CONTROL relay K1 remove the grid of V7B from the THRESHOLD potentiometer R45 and connect the grid to the manual gain control.

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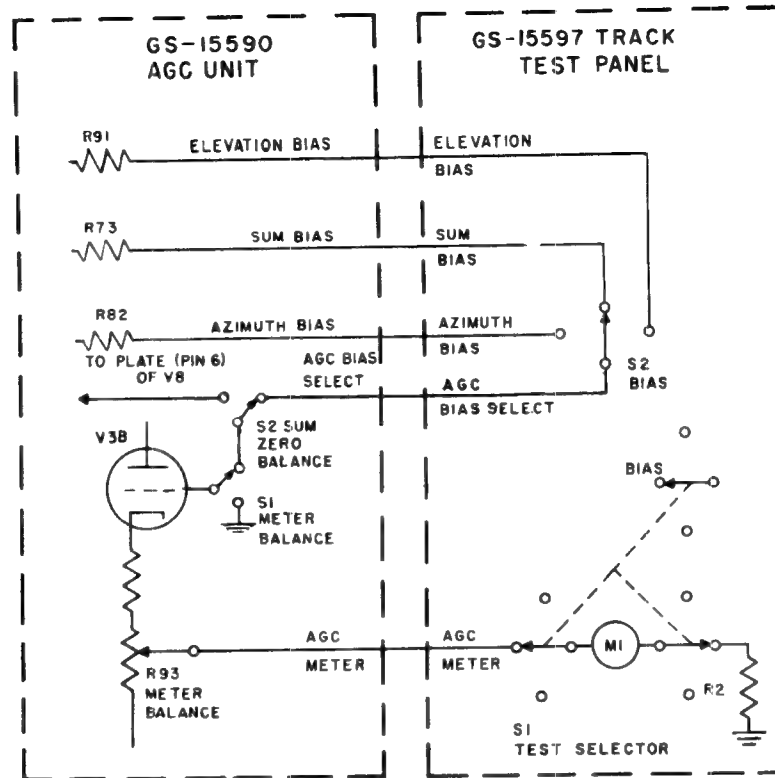
Figure 34. AGC metering circuit, simplified schematic.

(a)   When the manual gain control is used from the track test panel, spring-loaded GAIN CONTROL switch S5 is pushed and held in the MANUAL position. Contacts 5 and 6 insert MANUAL GAIN potentiometer R3 into the grid circuit of tube V7B in the AGC unit. Varying this potentiometer varies the bias applied to grid 6 of V7B in the AGC unit and thus the output of the AGC unit. TEST relay K1 in the track test panel is energized by placing TEST-OPERATE switch S12 in the TEST position in the target track control drawer. MAN GAIN CONTROL relay K1 has to be energized for this circuit operation.

(b)   When manual gain control is used from the target control panel, AGC-MGC switch S7 is pushed to the MGC position. This action energizes MAN GAIN CONTROL relay K1 in the AGC unit, provided TEST-OPERATE switch S12 in the target track control drawer is in the TEST position. TEST relay K1 in the track test panel is energized inserting GAIN potentiometer R14 in the grid circuit of tube V7B of the AGC unit. GAIN potentiometer R14 is then used to change the AGC output voltage.

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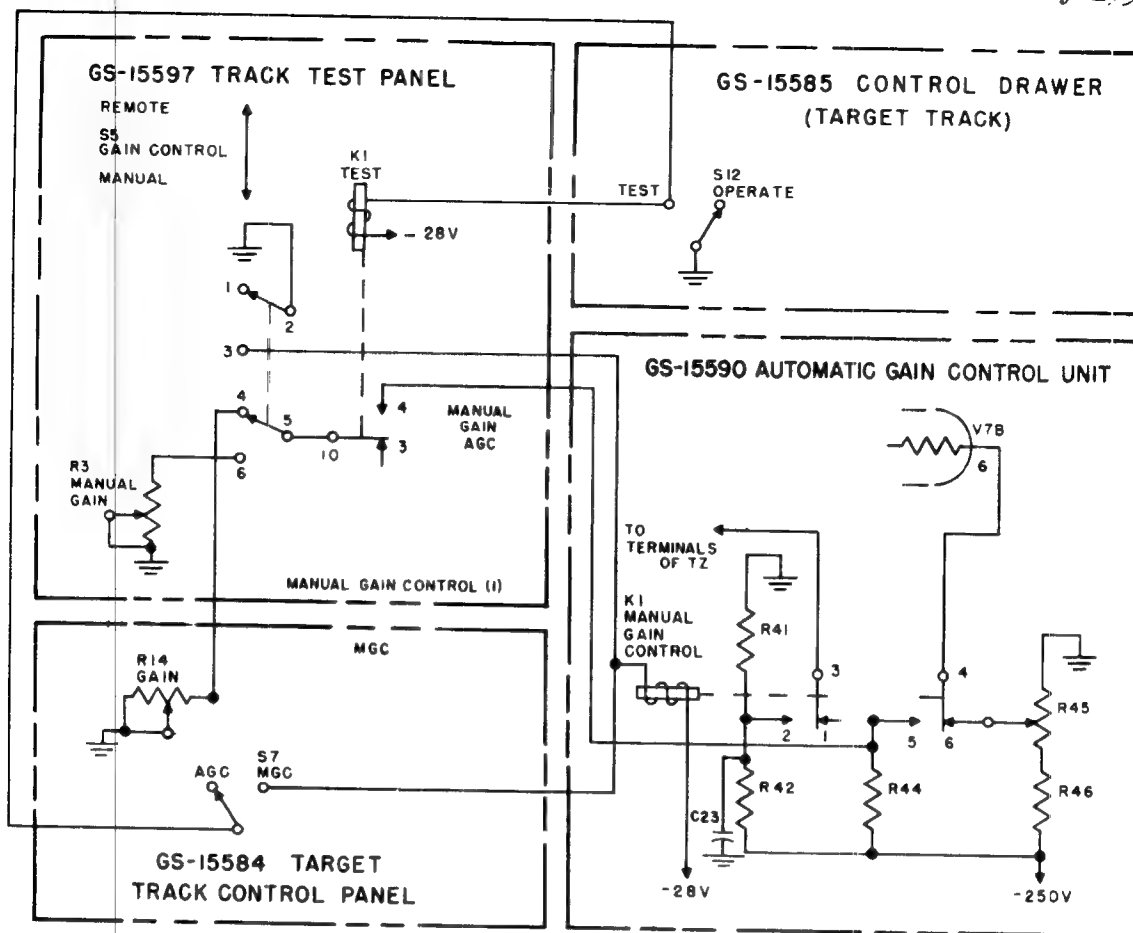
Figure 35.   Manual gain control, simplified schematic.

29.   SLIPRING SIGNAL TRANSFER

Figure 3-7 is the schematic of slipring assembly GS-15536.   These sliprings transfer the three i-f signals (sum, elevation and azimuth) from the movable part to the stationary part of the antenna.   Sliprings 1, 2, and 3 are used to transfer the sum i-f signal, rings 5, 6, and 7, the azimuth i-f signal, and rings 9, 10, and 11, the elevation i-f signal.   To reduce greatly the cross-talk between the three channels, the signal rings (2, 6, and 10) are triple-spaced from the ground rings on each side of the signal rings.   To reduce cross-talk further, a ring has been inserted between two adjacent grounds having a function not affecting the i-f signals (rings 4 and 8).

a.   Impedance matching.   The coaxial leads carrying the signals from the i-f preamplifiers are fed into these sliprings with a 50-ohm impedance.   This matches the impedance of the 250-foot coaxial cable which carries the three signals to the main i-f amplifiers.   However, the sliprings themselves present

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a capacitance of about 70 micromicrofarads to the circuit, which corresponds to about 38 ohms capacitive reactance at 60 megacycles. Since the coaxial cables are about 50-ohm lines, the sliprings have to be matched to the cables. Across each of the cables entering the sliprings (from the i-f preamplifiers) is connected a coaxial stub of the proper length that supplies the inductance required to resonate the slipring capacitance. Thus complete matching of impedances throughout the transmission of i-f signals from preamplifiers to main amplifiers is accomplished.

b. Phase shift. Rotation of the sliprings changes the length of the path over which a signal has to travel in going through sliprings. Since monopulse theory is based on phase shifts of received signals, the sliprings will consequently give an undesired shift of phase in the i-f signals. To combat this, the contacts (inputs and outputs) of the three signals are arranged symmetrically, thus giving all three signals the same amount of phase shift at the same time. Phase shifts, therefore, can be of any amount, as long as all three signals have the same amount of shift.

30. TRIPLE COAXIAL CABLE SIGNAL TRANSFER

The GS-51186 cable consists of three RG-9/U type coaxial conductors. The flexible outer jacket insulation is thin but able to withstand all-weather rough treatment. The over-all diameter is one inch, the length, 250 feet. This cable run 38 is connected between the main junction box on the target antenna trailer and a junction box on the radar control trailer. The main requirement of this cable is that there be no more than 10° of electrical phase shift between any two signals in the cable, regardless of temperature changes. There is approximately a 4-db loss of power for a signal passing through any one of the three coaxial conductors.

31. I-F MAIN AMPLIFIER

Three track i-f amplifier units GS-15589 are used in the TTR receiving system. These units are located in radar range and receiver cabinet assembly GS-15516. (For the physical location of the units, see sheet 24 of section XX of TM 9-5000-25.)
Signals from the i-f preamplifiers reach the i-f main amplifiers through the 250-foot triple coaxial cable. Each main amplifier has seven stages of amplification, a grounded-grid triode input stage and six succeeding pentode-amplification stages. The sum, azimuth error, and elevation error channels each require one of these identical i-f main amplifier units.

a. Block diagram discussion. Since azimuth, sum, and elevation i-f main amplifiers are identical in their operation, only the azimuth i-f main amplifier will be discussed. The other two amplifiers operate with their respective signal inputs.

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(1) Input network. The input network has two jacks and an array of resistors to give the proper input. When the TEST-OPERATE switch is in the TEST position, a 60-mc signal from the i-f test panel is coupled to the circuit for testing this unit and the AGC unit. Otherwise, only the 60-mc signal from the i-f preamplifers are coupled in.

(2) Grounded-grid amplifier. Refer to paragraph 25a for a discussion on the grounded-grid amplifier.

(3) I-F amplifiers V2 through V6. Amplifiers V2 through V6 are conventional amplifiers with an AGC bias furnished by the AGC unit. These amplifiers are in cascade and the coupling transformers are approximately 10-mc bandpass transformers.

(4) Power amplifier. Tube V7 is the power amplifier and does not incorporate AGC. The over-all gain of the i-f sum amplifier is approximately 110 decibels.

b. Detail circuit discussion.

(1) Input stage. Refer to figure 6-12. A JAN 5842 grounded-grid triode is used for the input stage to the unit. This tube gives low amplification, but has a good signal-to-noise ratio. Heavy loading of the output circuit eliminates the triode's natural tendency to break into oscillations and maintains the broad-band response necessary for good reproduction of pulses. The input signal from the 250-foot coaxial cable is applied through IF INPUT jack J2, resistor R36, and capacitor C1 to the cathode of the triode. The input impedance of the stage is low and consists of a voltage divider to match the impedance of the cable. Thus the noise level of the cable is reduced and undesired echoes in the cable are attenuated. The impedance of the input is equal to one over the mutual conductance $\left(\dfrac{1}{gm}\right)$ of the triode plus a function of the load impedance. Resistor R1 is bypassed by capacitor C2 and does not effect the input impedance directly. R1 establishes the operating point of the triode and therefore, its mutual conductance (gm) and the input impedance. Capacitor C4 is used to adjust the output impedance of V1 to match the impedance of interstage transformer T1 for 60-mc operation. Resistor R3 is shunted across the primary of transformer T1 to obtain the desired 10-mc bandpass for the transformer.

(2) Amplification. Stages V2 through V7 use JAN 5847 pentodes which have a nominal mutual conductance of 12,500. Interstage coupling is done with air core transformers T1 through T6, all of which cover approximately 10 megacycles. Automatic gain control (AGC), used on stages 2 through

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6, is applied through pin 1 of plug P1 from automatic gain control unit GS-15590. Decoupling networks Z8 through Z12, together with the associated r-f bypass capacitors C21 through C26, prevent undesired interaction between the stages through the common AGC input. Resistors R6, R11, R16, R21, R26, and R31 are bypassed by capacitors C5 through C10, respectively, to obtain an optimum operating bias for the associated stages. Because of the basic operation of a monopulse system, phase variations have to be the same for the three i-f channels. Because phase variations occur with gain variations of these stages, the following reasoning produced the circuit for minimum phase variations with gain changes. The unbypassed cathode resistors R5, R10, R15, R20, R25, and R30 are used for the purpose of disposing of variations in the input capacitance to these stages. Because the input capacitance increases with gain, the capacitive susceptance increases with gain. However, the feedback from these resistors also increases with gain, thus producing an inductive susceptance. This inductive susceptance acts to cancel the undesired input capacitive susceptance and thus minimizes the phase variations due to gain. The time constant of the cathode circuit is 0.04 microsecond. This permits rapid decay of an overload bias due to a strong signal. Resistor R34 and networks Z2 through Z7, in combination with capacitors C13 through C20, provide the necessary decoupling to avoid interstage action due to common plate and screen voltage supply. Filtering networks L2 through L9 and capacitors C27 through C35 prevent interaction between stages through filament supply. Tube V7 is the output stage and does not incorporate automatic gain control. Fixed capacitor C11 and variable capacitor C12 are used for matching the output of transformer T7 with the output coaxial cable. This cable conducts the i-f signal from IF OUT jack J3 to the video and phase unit. The optimum output is a 60-mc, 2-volt, peak-to-peak signal, fed into the 70-ohm coaxial cable. Over-all gain of this unit is approximately 111 decibels. Jack J4 is used to apply the i-f signals from i-f test panel GS-15608 for testing this unit and the AGC unit.

32. VIDEO AND PHASE UNIT

The three received signals (sum, azimuth error, and elevation error) in i-f form travel in three identical, but separate paths. In their paths from the antenna waveguide plumbing, through the i-f amplifiers, sliprings, cables, etc., these signals undergo approximately 15,000° of phase shift apiece. The inherent properties of tubes, resistors, and the various components which make up these identical paths are slightly different. Therefore each of the three signals will experience a differenc amount of phase shift that is detrimental to monopulse radar. To compensate for this video and phase unit GS-15677 makes the phase shift commonly equal to all three channels from the output of the r-f and antenna system to the input of the angle error detectors in the receiving system. The video and phase unit does this by providing an adjustable phase shift of

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approximately ±80° in each channel. In addition to this, the unit provides meter indication of the level of the IF signal of each channel, control information (video pulses) to the AGC unit, and received signal information to the missile AFC unit (the missile AFC unit operates in the target radar, though its output is not used during normal operation). The video and phase unit is located in radar range and receiver cabinet assembly GS-15516. (For the location of this unit, see sheet 24 of section XX of TM 9-5000-25.)

a. Block Diagram Discussion (fig. 6-15). The sum, azimuth error, and elevation error 60-mc, IF signals from the IF main amplifiers enter through SUM IN jack J3, AZ IN jack J1, and EL IN jack J7, respectively. Each signal enters a separate block labeled paraphase amplifier. The signals are coupled to two stages of IF amplification through a phase adjust and balancing network. After phasing and amplification, the output of the 60-mc signals in the elevation channel and the azimuth channel, and one of the two 60-mc signal outputs in the sum channel are applied through EL OUT jack J8, AZ OUT jack J2, AZ SUM OUT jack J5, and EL SUM OUT jack J6, respectively, to the angle error detector GS-15718 (elevation and azimuth). A portion of the output in each of the three channels is detected by crystals and applied to their respective cathode followers. From these cathode followers, the three signals are connected to a pulse-level metering circuit. This circuit consists of a signal selecting circuit, a pulse amplifier, a gate amplifier, a gated pulse-level detector, and an output cathode follower. The gating action is applied to this circuit from range error detector GS-15593 through RCBR GATE in jack J14. The output of the pulse-level metering circuit is applied to a meter in the track test panel. Part of the output of the sum cathode follower is applied through a video amplifier and out through jack J10 to the range error detector. This 0.3-μ sec pulse is the video information for the ranging circuits in the ranging system. Another part of the output of the sum cathode follower is applied through AGC PULSE OUT jack J11 to the AGC unit.

   (1) (Superseded) Paraphase amplifiers V1, V4, and V7. Since the three stages are identical, only one stage will be discussed. Two outputs are taken from V1, one from its plate and another from its cathode. Since the plate load and cathode resistors are of identical value, the cathode output will be identical in amplitude and waveshape and 180° out of phase with the plate signal. After passing through the phasing circuit, the output signal is 90° ± 80° out of phase with the input to the paraphase amplifiers, depending upon the adjustment of the phasing network. This adjustment is necessary to compensate for unequal phase shift between the sum and error signals through the receiving system so that the proper phase relation is provided in the angle error detectors.

   (2) (Superseded) IF amplifiers V2, V3, V5, V6, V8, and V9. In acquiring the desired phase shift of the signal, loss in signal strength is sacrificed. Therefore, more if amplification is provided by V2 and V3. These outputs are coupled to their respective angle error detectors as 60-mc, 0.2-μ sec pulses which vary in amplitude, depending upon the target antenna pointing error. A portion of the outputs is coupled to cathode followers through detectors and pulse stretchers.

   (3) Detector and pulse stretcher crystals CR1, CR3, and CR4. Detector and pulse stretcher crystals are used to produce only negative pulses that are 0.3 microsecond in duration and 2 volts in amplitude for the ranging system and to indicators on the IF test panel. One output from jack J11 is used in the AGC unit as the AGC video signal. These signals, before coupling to the output jacks, go through cathode followers.

   (4) Cathode followers V10A, V13A, and V15A. These cathode followers are used as impedance matching devices between the detectors and the metering equipment in the IF test panel.

(5) *IF signal level metering circuit.* The IF signal level metering circuit is composed of tubes V10B, V12A, V12B, V13B, and V15B and their associated circuitry. Together with the input signals from the various sources, an IF signal level indication is obtained on the LEVEL meter in the IF test panel. Tube V15B, a conventional cathode follower, couples the 15-volt, 0.4-μ sec pulse from the range error detector to the pulse level detector V12. Tube V13B is a pulse amplifier that amplifies and inverts the pulse from the cathode followers. The crystals in the input circuit to V13B are used to channel only one signal to V13B during testing. Meter cathode follower V10B is used to match impedances and isolate the previous stage from the remainder of the pulse level metering circuit. Tube V12, a pulse level detector, acts with capacitor C56 as an integrator capacitor that stores the charge of the dc level so as to control the conduction of the meter cathode follower. LEVEL ZERO potentiometer R94 is used to adjust the charge on capacitor C56; LEVEL SENSITIVITY potentiometer R85 is adjusted to give full-scale deflection of TEST meter M1 for a 1-volt input signal.

(6) *Range video amplification.* Tube V11, connected as a conventional video amplifier and cathode follower, amplifies and couples the pulse to the range error detector.

b. Detail Circuit Discussion (fig. 6–16).

(1) *Phase shifter, amplifier, and detector.*

(a) The 60-mc azimuth IF signal from track IF amplifier GS–15589 (azimuth) enters AZ IN jack J1 and is applied to the control grid of tube V1. Resistor R3 and inductance L1 are used to match the impedance of the 70-ohm coaxial cable connected at jack J1. Since plate load resistor R2 and cathode resistor R4 have identical values, the outputs taken from V1, one from its plate and another from its cathode, will be of equal amplitude but 180° out of phase. The plate-to-ground capacitance formed by interelectrode capacitance of the tube and capacitor C64 is balanced with the cathode-to-ground capacitance by

adjustment of PH BAL AZ capacitor C2. Also the coupling cathode resistor R5 has been selected to give the resistive elements of the circuit an ohmic value identical to the capacitive reactance of plate coupling capacitor C3 at its midcapacitance for 60-mc operation. Capacitor C72 is a d-c blocking capacitor which isolates the cathode potential from the R-C phasing circuit. Figure 36 is a simplified schematic of the phase-shifting network. Figure 37 is a vector diagram of the phasing voltages. For the sake of simplicity, capacitor C3 and resistor R5 are taken as the R-C phasing circuit. These points labeled on figure 36 correspond to like points on figure 37. Since the three comparison voltages (E_1 from cathode, E_2 from plate, E_0 from the combined output) are all referred to ground in the schematic, their respective vectors are all referred to common point A. When an alternating voltage is applied across a series R-C circuit, the voltage across the capacitor leads the voltage across the resistance by 90°. Since (in figure 37) the vectors *lags*

$$\overline{E_{CB}} + \overline{E_{BD}} = \overline{E_{CD}} \qquad \qquad (2)$$

lags

the vectors $\overline{CB}$, $\overline{BD}$, and $\overline{CD}$ form a right triangle with the right angle vertex at point B. Consequently, point B must be on a semicircle with AC ($E_1 + E_2$) as the diameter, this being the locus of all points so that $\overline{CB} + \overline{BD} = \overline{CD}$, and $\overline{CB}$ (Ec) leads $\overline{BD}$ (Er) by a right angle. The position of point B on the semicircle depends on the ratio of capacitive reactance (X_{C3}) to resistance (R_5). The capacitance (C3) is variable; the resistance (R5) is fixed. When capacitor C3 is adjusted so that its adjusted capacitive reactance at 60 megacycles equals the resistance of R5, vector $\overline{CB}$ and $\overline{BD}$ have equal amplitude, and voltage E_0 $\overline{(BA)}$ experiences a 90° phase difference from voltage E_1. Therefore:

$$\text{Tan } \theta = \frac{R}{X_C} \cdot \qquad \qquad (3)$$

When capacitor C3 is adjusted to increase its capacitive reactance, the magnitude of vector $\overline{CB}$ increases and the magnitude of vector $\overline{BD}$ decreases. Therefore, voltage E_0 ($\underline{/\emptyset}$) experiences less than 90° of phase shift with respect to E_1. Conversely, if the capacitive reactance of capacitor C3 decreases, the magnitude of $\overline{CB}$ decreases and the magnitude of $\overline{BD}$ increases, and E_0 ($\underline{/\emptyset}$) experiences more than 90° of phase shift with respect to E_1. The average limits of phase shifting by adjusting capacitor C3 is a 10° to 160° phase lag of E_0 ($\underline{/\emptyset}$) with respect to E_1. Considering the case where the capacitive reactance of capacitor C3 equals the resistance of the circuit, the output voltage E_0 actually lags the input voltage on pin 1 of V1 by 90°. Considering, however, the output voltage (E_0) in this particular case as being the

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0° phase reference, the output voltage can be effectively varied approximately ±80°. After having determined angle θ by use of equation (3), angle Δ is its complement. A resume may well be appropriate at this point. Angle θ is known from equation (3), E_2 and E_1 are known from the value of the resistors in the cathode and plate circuit of V1 to be equal in amplitude but 180° out of phase, with point A in the vector diagram as ground. E_o is known to be the radius of a semicircle, or equal to E_1 or E_2. Then from the above information $\overline{BA}$, $\overline{AD}$, and angle Δ is known. With this information $\underline{/\emptyset}$ is determined by the law of sines:

$$\text{Sin}\emptyset = \frac{BD\sin\Delta.}{BA} \qquad (4)$$
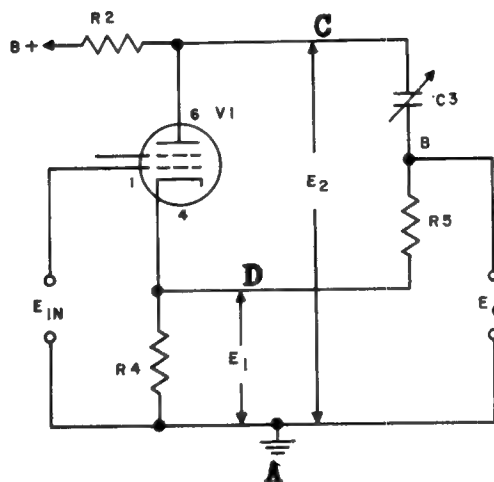
Figure 36. Phase shifting network, simplified schematic.
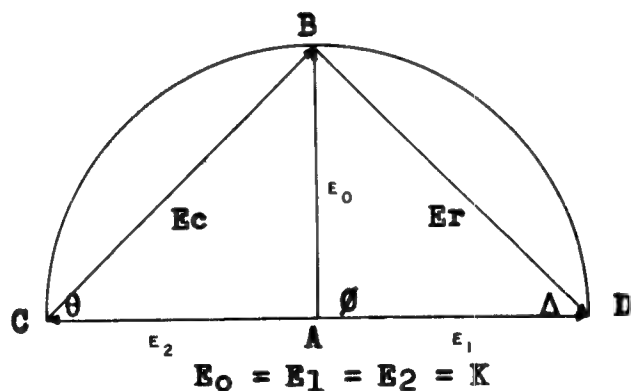

$$E_o = E_1 = E_2 = K$$

Figure 37. Phase shifter vector diagram.

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(b) Capacitor C2 is adjusted at the factory to make the cathode-to-ground capacitance of tube V1 equal to the plate-to-ground capacitance. Inductance L2 is used to compensate for the input inductance to the tube and does not affect operation of the phasing circuit. If more than $\pm 80°$ of phase shift is required for optimum operation, a piece (or pieces) of RG/59U coaxial cable may be patched between the cable from the i-f main amplifier and input jack J1. A 31-inch length of this cable provides approximately +90° of phase shift in the signal.

(c) The signal from the phase-shifting circuit is developed across resistor R8 and applied to the control grid of V2. The signal is amplified and applied through transformer T1 to the control gird of V3. The bandpass of T1 is made 10 megacycles by the shunting action of resistors R7 and R11. This is desirable since automatic frequency control does not maintain an exact 60-mc i-f. Capacitor C5 in the cathode circuit of V2 bypasses most of the cathode resistance in order to have cathode bias. However, unbypassed resistor R9 provides a small amount of degeneration which effectively stabilizes the control grid-to-ground capacitance of V2. Though tube V3 is an amplifier identical to V2, the primary of transformer T2 has an 8-mc bandpass. Impedance networks Z1 and Z2 are used to prevent interaction of the 60-mc signals between the plates of tubes V1, V2, and V3. The 60-mc output from transformer T2 is partly applied through AZ OUT jack J2 to the azimuth angle error detector. Resistors R16, R17, and R18 form a voltage divider network to give an output of the azimuth i-f of the desired amplitude with respect to the i-f sum channel amplitude. This is necessary since the angle error detector requires a definite ratio of the amplitudes of the azimuth and sum i-f signals, and the sum output in the video and phase unit is automatically reduced because of the other sum channel outputs. Part of the 60-mc signal from T2 is applied to crystal CR1. This crystal detects the signal by conducting only during the negative half-cycles. Inductance L3 and capacitor C10 filter the rectified signal output from CR1 and effectively eliminate the 60-mc variations to produce a 0.3μsec video pulse. This pulse is developed across resistor R57 and is applied through parasitic suppressor resistor R61 to the grid of cathode follower V10A.

(d) The circuitry of the elevation i-f channel in phase-shifting, amplification, and detection is identical to that of the azimuth channel discussed above. The 60-mc elevation i-f signal from the (elevation) track i-f amplifier GS-15589 is applied to the control grid of tube V7 through EL IN jack J7. The output of the 60-mc i-f is applied through EL OUT jack J8 to the elevation angle error detector. A portion of the output signal from transformer T6 is detected by crystal CR4 and, as a result, negative 0.3μsec video pulses appear on the control grid of cathode follower V15A.

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(e)   The 60-mc sum i-f signal from the (sum) track i-f amplifier GS-15589 enters SUM IN jack J3. A portion of this signal is applied through resistor R21 and AFC OUT jack J4 to the missile automatic frequency control unit GS-15592. The missile AFC unit functions in the target radar, but its output is not used since the (target) AFC unit controls local oscillator frequency. Circuitry from input jack J3 to output of transformer T4 is identical to the azimuth channel previously discussed. Part of the 60-mc output from transformer T4 is applied through AZ SUM OUT jack J5 and EL SUM OUT jack J6 to the azimuth and elevation angle error detectors, respectively. The remainder of the 60-mc output from transformer T4 is detected by crystal CR3 and applied through inductance L12 and resistor R68 to the grid of cathode follower V13A as a negative 0.3μsec video pulse.

(f)   The 0.3μsec negative video pulses from the azimuth, sum, and elevation channels are presented to cathode followers V10A, V13A, and V15A. These three cathode followers are identical in circuitry. Fixed bias is supplied by a voltage divider, a 25K resistor and the cathode resistor of each tube, connected between -250 volts and ground. The outputs from these cathode followers at AZ ERROR PULSE jack J9, SUM PULSE jack J12, and EL ERROR PULSE jack J13 are for oscilloscope-monitoring purposes. These are connected only when this unit is being tested. The output of tube V13A, connected through capacitor C48 to AGC PULSE jack J11, is applied to the automatic gain control unit GS-15590. This signal is used as the control video with which the gain of the i-f main amplifiers is controlled. The outputs of the cathode followers in the three through-coupling capacitors C49, C60, and C61 are applied to the the signal level metering circuit, discusses in paragraph (3) below.

(2)   Range video amplification. Part of the output from cathode follower V13A is applied through coupling capacitor C71 to the grid of video amplifier V11A. Tubes V11A and V11B are connected as conventional R-C coupled video amplifiers. Fixed bias on the grid of V11B is obtained from the voltage divider between -250 volts to ground consisting of resistors R66 and R67. Tube V11B is connected as a cathode follower with the sum video output applied through RANGE VIDEO jack J10 to range error detector GS-15593 in the ranging system. This output consists of positive 0.3μsec video pulses of approximately 2 volts amplitude. Crystal CR8 in the grid circuit of V11B prevents any negative signals from being applied to the range error detector from the output of V11B.

(3)   I-F signal level metering circuit.

(a)   Tubes V10B, V12A, V12B, V13B, and B15B, with their associated circuit components, provide a means for indicating with a meter the level of

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strength of any one of the three i-f signals (fig 38). This is a simplified schematic of the circuit which selects one of the three signals to be monitored and the metering circuit. If the azimuth signal is desired to be monitored, LEVEL switch S3 in track test panel GS-15597 is turned to the AZ position. This action places ground on the lower side of resistor R95 and crystal CR11 in the video and phase unit and permits conduction of the azimuth video pulses through crystal CR6. Without crystal CR11 connected in the circuit, video variation across capacitor C60 would be clipped by crystal CR6 since these variations would fluctuate above and below ground potential. However, with crystal CR11 connected in parallel with resistor R95, the video pulses are clamped below ground potential and pass to crystal CR6 as negative signals. When a signal other than the azimuth signal is being monitored, LEVEL switch S3 in the test panel removes ground from the lower side of resistor R95 and crystal CR11. The voltage divider, consisting of resistors R97 and R98 connected from +150 volts to ground, then applies approximately +10 volts to the cathode side of crystal CR6. The amplitude of the azimuth signal will have an average somewhat less than -1 volt, and never exceeding -10 volts. Therefore, crystal CR6 is not able to conduct during the application of the azimuth signal and this signal does not appear on the grid of tube V13B. The selection of the other two signals is similar to this, with the result that only one signal is applied to V13B at one time.

(b) The +15-volt 0.4μsec receiver gate pulse from the range error detector GS-15593 is applied through RCVR GATE jack J14 to the grid of cathode follower V15B. The gate is then applied through capacitor C70 to the two primary windings of transformer T7.

(c) The two sections of tube V12 and their circuit components make up a pulse level detector. This is a gating circuit used to permit only the desired pulse to be monitored. Since no gating has taken place in the paths of the three i-f signals up to and through this unit, returned signals from objects other than the target being tracked may enter this video and phase unit. To observe the level or strength of the desired pulses only, the gating circuit is necessary.

(d) The positive gating pulse from tube V15B applied to the primaries of transformer T7 causes positive pulses to appear on the grids with respect to the cathodes of tubes V12A and V12B. Thus these two tubes are permitted to conduct, provided that potentials are appropriately applied to their cathodes and plates. When the gating pulse is removed from the primary of transformer T7, the charges built up on capacitors C57 and C58 cause current to flow through resistors R90 and R91. This action applies a negative voltage on the grid with respect to the cathode of each tube, which is sufficient to hold these tubes cut off until the next gating pulse is applied.
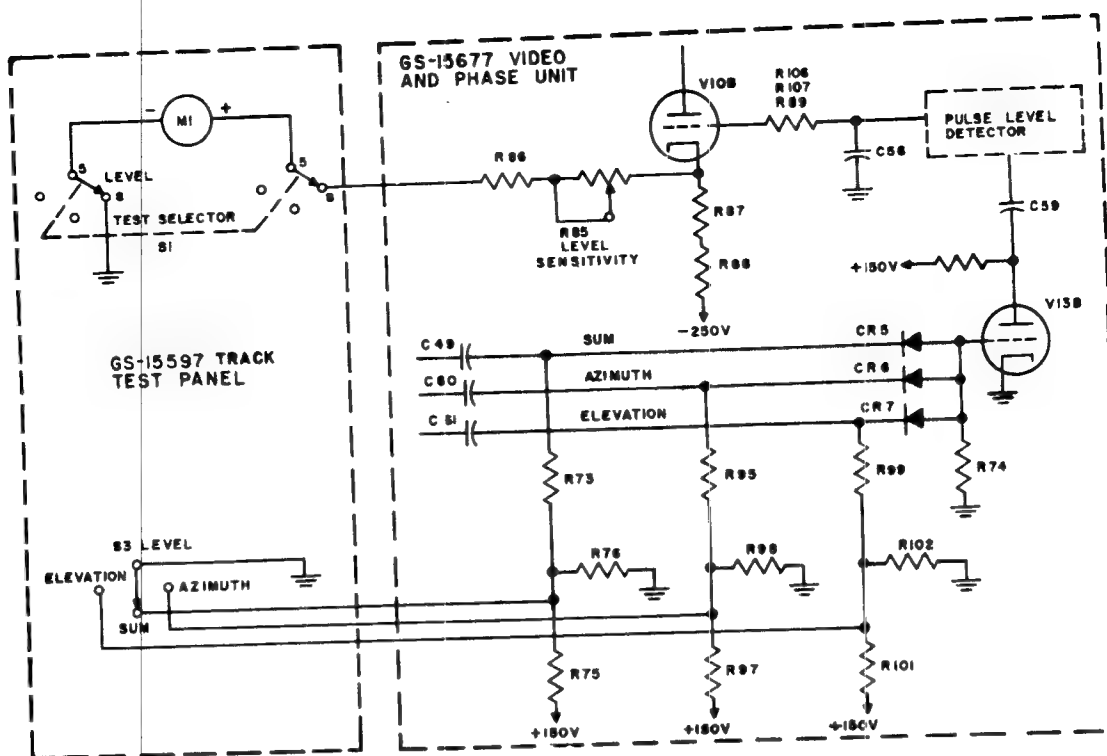
Figure 38. I-F level monitoring circuit, simplified schematic.

(e) The positive output of tube V13B is applied through capacitor C59 and crystal CR13 to tubes V12A and V12B to obtain a monitoring indication of the i-f pulse amplitude. Video variations appearing across capacitor C59 are clamped above a fixed d-c potential determined by the setting of LEVEL ZERO potentiometer R94. The clamping circuit is formed by capacitors C59 and C77, potentiometer R94, and crystals CR9 and CR13. This circuit permits the full amplitude of the video signal to pass through CR13 virtually unclipped. The amplitude of these video pulses changes with the strength of the i-f input signal. Consider the case where a pulse of amplitude A is applied from CR13 to the cathode of tube V12A and the plate of V12B. At this same time the gating pulse is applied to V12A and V12B allowing them to conduct. However only V12B conducts, since the positive level pulse is applied to its plate and to the cathode of V12A. Conduction of V12B charges capacitor C56. When the charge on capacitor C56 reaches amplitude A, V12B ceases conducting since its plate and cathode reach the same potential. During this 0.4μsec duration, V12A is not conducting because its plate is negative with respect to its cathode. When the gating pulse is removed and both V12A and V12B are cut off, capacitor C56 has no d-c path to ground through which to discharge. Therefore this charge of amplitude A is retained by C56 until the next gating and received pulses are applied. If the next video pulse and several of the succeeding pulses to crystal

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CR13 have an amplitude B which is less than amplitude A, it would appear that as the gate is applied capacitor C56 would retain the amplitude A charge because of the back resistance of crystal CR13. However the back resistance of CR13, although quite high, permits sufficient conduction through V12A to charge capacitor C56 to amplitude B. When the charge of capacitor C56 reduces to amplitude B, the plate and cathode of V12A have the same potential and the tube ceases to conduct. Thus after several gating pulses, a new charge of amplitude B is obtained on capacitor C56, making it act as an integrator capacitor. As described above this action seems sequential and slow, but it actually happens within a period of a few milliseconds. When tube V12B conducts, capacitor C56 charges positively, when tube V12A conducts, capacitor C56 charges in a negative direction.

(f) Capacitor C56 retains its charge for approximately 1,000 microseconds between the gating pulses. The amplitude charge on capacitor C56 is applied to the grid of tube V10B. Therefore as the video-pulse level through capacitor C59 and crystal CR13 changes, the charge on capacitor C56 changes, and current flow through tube V10B changes in the same proportion. Refer to figure 38. Meter M1 in the test panel GS-15597 is connected to the cathode of tube V10B, provided TEST SELECTOR switch S1 in the same test panel is in the LEVEL position. In this manner the cathode potential of tube V10B appears on LEVEL meter M1 giving an indication of the level of the video pulses.

(g) LEVEL ZERO potentiometer R94 is used to adjust the charge on capacitor C56 with zero signal input, to obtain a zero reading on meter M1 in the test panel. Potentiometer R94 applies a d-c potential through crystals CR9 and CR13 in a reverse direction. With no signal input this d-c potential is adjusted to cause the cathode of V10B to be at ground potential. LEVEL SENSITIVITY potentiometer R85 is adjusted to give proper deflection of LEVEL meter M1 for a 1-volt input signal.

33. AZIMUTH AND ELEVATION ANGLE ERROR DETECTOR (MODIFIED)
(fig 6-20.1)

The received azimuth error i-f signal and a portion of the received sum i-f signal from the video and phase unit are applied to angle error detector GS-15718. This unit uses these signals to produce a video pulse signal representing the direction and magnitude with which the antenna is pointing off target in azimuth. In ideal radar operation if the antenna is off target in azimuth in one direction, the azimuth error i-f signal will be 180° out of phase with the sum i-f signal. If the antenna is off target in the opposite direction, the two i-f signals will be in phase. The angle error detector compares the phase relationship of the two signals and produces negative or positive video pulses, depending on the direction the antenna is off target in azimuth. The amplitude of these output video pulses

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is proportional to the amount of azimuth error. If the antenna is pointing on target in azimuth, no azimuth angle i-f signal is applied to this unit because of r-f cancellation in the waveguide. This results in no output from the angle error detector. As the target moves in one direction increasingly away from the antenna azimuth on target position, the output of this unit will be negative video pulses of increasing amplitude. If the target moves in the opposite direction in the same manner, the output of the unit will be positive video pulses of increasing amplitude. The 3μsec expansion pulse from the ranging system is also applied to the angle error detector. This signal is used to gate the output video pulses, which insures that the only received i-f signals used are from the target being tracked. A portion of the output pulses from the azimuth angle error detector is applied to the presentation system. Here the pulses are used in pip canceling on the azimuth indicator. The elevation channel of the radar also employs an angle error detector which is used in the same way as in the azimuth channel just discussed. The two angle error detectors are located in the radar range and receiver cabinet assembly GS-15514 (see sheet 24 of Section XX TM 9-5000-25).

a. Block diagram discussion (fig 6-19).

(1) I-F amplifiers V1 and V2. Tubes V1 and V2 are conventional amplifiers. The input signal to V1 is coupled from the video and phase unit as the angle error signal, the input to V2 is coupled from the video and phase unit as the sum i-f signal used as a reference signal for phase and amplitude comparison. Resistor R41 is in the grid circuits of both tubes for gain control.

(2) Detector. Transformer T3A, tube V3, and associated circuitry make up the detector. The sum signal is coupled to the cathodes of V3A and V3B. The error i-f signals are applied to the plates of tube halves V3A and V3B. Through the combination of T3A, V3, and associated circuitry, the detector stage detects and compares the degree of angle error.

(3) Differential amplifier V4. The purpose of the differential amplifier is to combine the two inputs from the detector to produce one video pulse signal. This is done by the controlling current through the common cathode resistor. The output signal will be the amplitude of the error signal.

(4) Output cathode follower V6. The two tube halves of V6 are connected in parallel and as conventional cathode followers with a common cathode resistance (a 1-megohm resistor in the error pulse rectifier).

(5) Expansion notch amplifier V8. V8 is a conventional amplifier that amplifies the expansion notch from the ranging system. The load for the

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expansion notch amplifier V8B is the 5-6 winding of transformer T4. The crystal CR1 eliminates any negative overshoot on the positive 20-volt, 3μsec output pulse.

(6) Gating and limiting circuits. Since tubes V5 and V7 and transformer T4 interact, the gating and limiting circuits are treated together. The aforementioned components form a gated limiter circuit. This circuit provides limiting of noise signals in the output circuit of differential amplifier V4 except during the 3μsec gate period. This period occurs in coincidence with the error video output pulse from V4. Thus, the desired error signals are permitted to pass V5, and any circuit noise signals at other times are virtually eliminated by the action of V5.

(7) Monitor cathode follower V8A. A portion of the output pulses from the cathodes of V6 are applied to the grid of V8A. Through V8A a good reproduction of the error video signal is applied through MONITOR OUT jack J5 and developed across a 75-ohm resistor in the video error signal unit. The video error signal unit is not a part of the angle error detector.

b. Detail circuit discussion (fig 6-22.1).

(1) I-F amplifiers. See figure 6-20.1. Tubes V1 and V2 and their associated circuitry form two i-f amplifier stages. These tubes amplify the two input i-f signals to a usable level for the following detector circuit. The peak-to-peak amplitude of the sum input i-f is about one volt in normal operation. The peak-to-peak amplitude of the error i-f is constantly variable, and the 0.5-volt amplitude shown on the schematic is arbitrary to show a comparison of amplitudes of signals in the following circuitry. For this particular value, the antenna would be about 20 angular mils off target. The impedance of the input coaxial cables is matched by termination resistors R1 and R43. The voltage divider, resistors R41 and R42 between -250 volts and ground, supplies a common fixed bias for both amplifiers, therefore controlling their gain. GAIN potentiometer R41 is field adjusted to give the proper output of this unit under specified test conditions. Networks Z1 and Z2 and capacitors C2, C3, C9, and C10 are filtering (decoupling) components used to prevent interaction between the input circuits of V1 and V2, and to prevent the 60-mc i-f front from entering the -250-volt supply. Inductors L1 and L2 help form the proper input impedance for the two tubes. The stages of these two amplifiers are very similar to those in the i-f main amplifier unit. The plate load impedance of V1 and V2 consists mainly of resistor R5 (primary of transformer T1) and resistor R47 (primary of transformer T2), plus the impedance also reflected through the transformers. This results in transformers T1 and T2, having a bandpass of about 12 megacycles. Networks Z3 and Z4 and

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capacitors C5, C8, C13, and C14 provide filtering of 60 megacycles from the +150-volt supply. Capacitor C34 and variable capacitor C16 determine the phase relationship of the sum signal with respect to the error signal. Although theoretically there should be no error signal present when the antenna is pointing on target, there is a small error signal applied to the unit. This signal should be 90° out of phase with the sum signal. Capacitor C16 is adjusted for this purpose. Capacitor C15 is added to the circuit to provide resonance at 60 megacycles. Capacitor C17 is used for the same purpose in the secondary of transformer T1.

(2) Detector. Transformer T3, tube V3, and associated circuitry make up the detector. The sum i-f input signal is coupled through capacitors C31 and C32 to the cathodes of duodiode V3. The error i-f input signals are applied to the diode sections of V3 for rectification. The degree of rectification depends upon the phase relationship of the sum and error signals and the relative amplitudes of these signals. With an error signal applied, the net result is two video signals, which appear at the grids of tube V4. Figure 39 is used in the discussion of the operation of the detector circuit. This figure illustrates roughly the signal relationships at the various circuit points designated.

(a) Transformer T3A is connected so that a positive signal on terminal 5 with respect to terminal 6 appears as a positive signal on terminal 7 and as a negative signal on terminal 9. Thus, the two i-f signals, 180° out of phase, are applied to the plates of V3. Capacitors C35 and C33 provide adjustment for balancing the amplitude of these signals. (This is a field adjustment.)

(b) Assume that the antenna is pointing off-target in a direction which causes the error and sum i-f signals at tube V3A to be in phase. The sum i-f signal from winding 3-4 of transformer T3 will always be larger in amplitude that the error i-f applied from winding 5-6 of transformer T3. Therefore during the first half-cycle the cathode is more positive than the plate, and no conduction takes place in V3A. In the negative half-cycle, tube V3A conducts slightly and tends to reduce the negative swing at the cathode. The resultant 60-mc signal at the cathode of V3A follows the full positive portion of the sum signal but does not have its full negative swing. At tube V3B the error and sum i-f signals are 180° out of phase. In the first half-cycle the plate of V3B is negative and the cathode is positive; thus tube V3B does not conduct. In the second half-cycle the plate tends to go positive and the cathode negative causing relatively heavy conduction through V3B. The net result is a 60-mc signal at the cathode of V3B with the full positive portion of the sum signal and the negative portion practically canceled so that the net signal goes only slightly negative. The distributed

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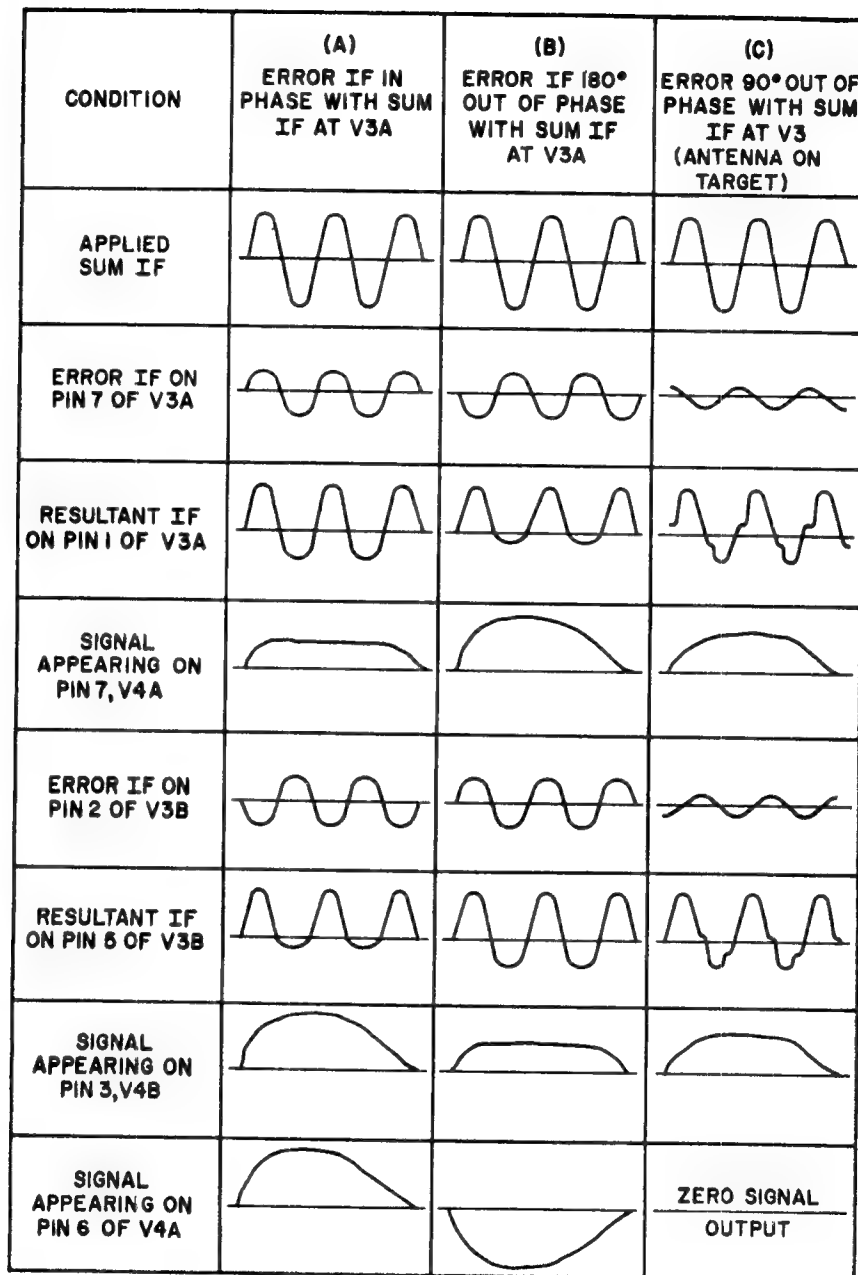| CONDITION	(A) ERROR IF IN PHASE WITH SUM IF AT V3A	(B) ERROR IF 180° OUT OF PHASE WITH SUM IF AT V3A	(C) ERROR 90° OUT OF PHASE WITH SUM IF AT V3 (ANTENNA ON TARGET)
APPLIED SUM IF			
ERROR IF ON PIN 7 OF V3A			
RESULTANT IF ON PIN 1 OF V3A			
SIGNAL APPEARING ON PIN 7, V4A			
ERROR IF ON PIN 2 OF V3B			
RESULTANT IF ON PIN 5 OF V3B			
SIGNAL APPEARING ON PIN 3, V4B			
SIGNAL APPEARING ON PIN 6 OF V4A			ZERO SIGNAL OUTPUT

Figure 39. Angle error detector GS-15718, idealized waveforms.

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capacitance of the circuit plus capacitor C17 and the inductance from L6 and L7 filter the 60-mc signal so that positive video pulses appear at the grids of tube V4. The amplitude of the positive video pulse at the grid of V4A is larger than the amplitude of the positive video pulse appearing at the grid of V4B for the condition under discussion. The difference in the amplitudes of the two signals causes the circuit of V4B to produce one positive video pulse. A variation in the amplitude of the error signal appearing across winding 5-6 of T3 produces proportional variation in the amplitude of the signal applied to V4. The result is a proportional variation in the amplitude of the positive video signal from V4B. The detector action in this condition is illustrated in A of figure 39.

(c) When the antenna points in the opposite direction off-target from that stated in subparagraph (b) above, the error i-f signal at terminal 5 of transformer T3 is 180° out of phase with the sum i-f signal at terminal 3 of transformer T3. This results in a larger positive video pulse at the grid of V4B than at the grid of V4A. In this case, the video output signal from V4B is negative and proportional in amplitude to the amplitude of the error i-f signal. The signal flow is represented by B in figure 39.

(d) For ideal operation, when the antenna points on-target there should be no error i-f signal presented to this unit. However, due to monopulse waveguide characteristics a small signal is presented to this unit for the on-target condition. This signal is 90° out of phase with the sum i-f signal. The adjustment for phase relationship is capacitor C16. In this on-target condition, assume the small signal at terminal 5 of transformer T3A lags by 90° the large signal at terminal 3 of transformer T3B. With these signals applied to V3, there is a small amount of conduction during the positive half-cycle of the applied sum signal. In the negative half-cycle, both tube halves of V3 increase conduction. The resultant signals at the cathode of V3 are shown in figure 39. These signals are filtered to produce video pulses at the grids of V4. In the condition just described, two video pulses, positive and equal in amplitude, are fed to the grids of V4. This produces no signal output from V4, representing no angle error. Waveforms of this condition are illustrated in C of figure 39.

(e) The cathode circuits of the two diodes have identical components. Capacitor C17 is field adjusted to make the capacitances of each circuit equal. Inductors L6 and L7 together with capacitor C17 provide equal i-f filtering in each circuit. MOD BAL potentiometer R9 is adjusted to cause equal conduction in the two diode sections of V3 for the same signal applied to each. The two output signals are applied through parasitic suppressor resistors R11 and R12 to the control grids of tubes V4B and V4A.

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(3) Differential amplifier. Tube V4 and associated circuitry make up the differential amplifier. The purpose of this circuit is to combine the two inputs from the detector to produce one video pulse signal. For circuit design the common cathode resistance for both sections of V4 is required to be large. Because these sections draw heavy current, resistors R16 through R19 are used for obtaining the proper resistance and for power dissipation. If positive signals of equal amplitude appear on the two grids of V4, increased conduction through V4A causes the cathode potential of V4B to rise. This potential rise is equal to the positive signal appearing on the grid of V4B. Effectively, this produces no change in conduction of V4B, and there is no signal output. This is the on-target condition. Assume now that the antenna is pointing off target, and the output of the detector circuit is a large positive pulse at the grid of V4B and a small positive pulse at the grid of V4A. Conduction through V4A is thus small and causes the cathode potential due to conduction of V4A to be just slightly positive. The positive signal on the grid of V4B causes the tube to conduct enough to cancel the effect of V4A and also to produce a negative video pulse at its place. The difference in the amplitudes of the input signals is amplified by this circuit and produced at the plate of V4B. Capacitors C18 and C20 provide a-c signal ground at their points of connection. Inductor L3 is a series peaking inductance used to improve the high-frequency response of V4. For signal inputs opposite to that discussed, the video signal output is positive. When the antenna is on-target no signal appears on the grids. The output of V4B is directly coupled to limiter V5 and output cathode follower V6. Because the action of V5 and its associated circuitry has no effect on the output of V4B at the time of each video pulse, V5 is not discussed at this point.

(4) Output cathode follower. The two sections of tube V6 are connected in parallel and used as a cathode follower for the output video pulses. The common cathode resistance for both sections is R22. The pulses are coupled through capacitor C27 and jack J3 to error pulse rectifier GS-15717. The termination for these pulses in the latter unit is a very small load (1-megohm resistor). The coaxial cable thus looks to this output cathode follower like a large capacitance. The tendency of this capacitance to act as a bypass capacitor across R22 is compensated for by the use of two tube sections. The negative pulses, therefore, cannot cut the tubes off.

(5) Expansion notch amplifier. The expansion notch from the ranging system is applied through jack J4, capacitor C21, and parasitic suppressor resistor R34 to the control grid of V8B. The 20-volt 3μsec negative pulse is amplified in conventional manner by V8B. The output positive pulse from V8B is applied to the primary of transformer T4 in the gating circuit. Crystal CR1 is used to eliminate any overshoot which might result from a tendency of transformer T4 to differentiate the gating pulse. Crystal CR1

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shorts any negative signals on terminal 6 with respect to terminal 5 of transformer T4. This short is also reflected to the other windings.

(6) Gating and limiting circuits. Tubes V5 and V7, transformer T4, and associated circuitry make up a gated limiter circuit. This circuit provides limiting of noise signals in the output circuit of differential amplifier V4 except during the 3μsec gating period. This time occurs in coincidence with the error video output pulse from V4. Thus, the desired error signals are permitted to pass V5 and any circuit noise signals at other times are practically eliminated by the action of V5.

(a) If the circuit is in the quiescent state, the voltage divider from ground through resistors R33, R29, R23, and R13 and inductor L3, plus the voltage divider action of the LEVEL LIMIT potentiometer R50 with resistors R48 and R49, and the conduction of V4 establishes the following potentials: +102 volts at pin 3 of V7A, +106 volts at pin 7 of V7B, and +110 volts at pin 6 of V4B. These voltages are not always this value, but their relationship to each other remains the same. The conduction of V7B with +106 volts on its grid is such to cause the voltage drop across the cathode resistor R26 to be 112 volts. By the same token, the voltage at the cathode of V7A is +108 volts. At pin 7 of V5A appears +108 volts and at pin 1 of V5A appears +110 volts. At pin 5 of V5B appears +112 volts and at pin 2 of the same section appears +110 volts. Any negative signals appearing at pin 6 of V4B which tend to cause the plate of V4B to become less than +108 volts are limited to +108 volts by the diode action of V5A. Tube V5B likewise limits any positive signals on the plate of V4B to +112 volts. Thus, noise signals are limited to a 4-volt peak-to-peak range at the plate of V4B. The over-all effect of this limiting action of V5A and V5B causes the average of these negative and positive noises to be 110 volts, or no signal.

(b) When the positive 3μsec gate from tube V8B is applied to transformer T4, terminal 6 becomes positive with respect to terminal 5. The induced pulse in the other two windings of transformer T4 causes terminal 1 to become negative with respect to terminal 2 and terminal 4 positive with respect to terminal 3. Therefore a positive pulse is applied through capacitor C25 to the grid of V7A. As a result, these two tubes produce 20-volt, 3μsec pulses on their cathodes with the same polarity as the pulses on their grids. Consequently the voltage at pin 7 of V5A during the gate is +88 volts and that at pin 5 of V5B is +132 volts. The limiter V5 now permits passage of any signals appearing at the plate of V4B having peak-to-peak amplitudes of 44 volts or less, varying around +110 volts. Therefore, all video error pulses are passed since their maximum amplitude should never exceed about 11 volts in the positive or negative direction. When the gating pulse is removed from transformer T4, the circuit returns to the 4-volt peak-to-peak limiting state. The limiting

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action of this circuit was designed to eliminate two sources of error: the main bang of the transmitter pulse and the ground clutter noise produced when the radar is pointing low in elevation. LEVEL LIMIT potentiometer R50 is adjusted to obtain optimum limiting operation of the circuit. This is a field adjustment.

(7) Monitor cathode follower. A portion of the output video pulses from the cathodes of V6A and V6B is applied through capacitor C26 and parasitic suppressor resistor R39 to the grid of V8A. The cathode resistance of V8A consists of resistors R37 and R38, and R51 to ground. R51 is paralleled by a 75-ohm resistor in the video error signal panel GS-15581. Therefore, a good reproduction of the error video pulses is applied through jack J5 and developed across the 75-ohm resistor. Resistor R51 in this unit is used for protective purposes in case the cable is removed from jack J5 during operation or testing.

34. ERROR PULSE RECTIFIER (MODIFIED) (fig 6-22.1)

Two error pulse rectifiers GS-15717 are used in the target-tracking radar, one for the elevation channel and one for the azimuth channel. The error pulse rectifiers accept the angle error video pulses from the respective angle error detectors and convert the information contained in these pulses into d-c signals. The d-c outputs of these units are applied to the respective angle modulator units GS-15527 in the antenna positioning system. When the antenna is pointing on-target, no angle error video pulses are applied to the error pulse rectifier. The output of the unit is a zero d-c signal with respect to ground. If the antenna is pointing off-target so that positive video signals are applied to the unit, a positive d-c signal with respect to ground will be the output. For negative input video pulses, there will be negative d-c signals out. The amplitude of the d-c output signals vary proportionally with the peak amplitude of the video signals applied. The two error pulse rectifiers are located in the radar range and receiver cabinet assembly.

a. Block diagram discussion (fig 6-21.1).

(1) Pulse amplifier V1. Tube V1 and associated circuitry comprise a con-- ventional pulse amplifier. The inverse feedback is used to improve the frequency response of the entire amplifier circuit, therefore reproducing the desired input pulse shape at the junction of resistor R27 and capacitor C15.

(2) Paraphase amplifier V2. Tube V2 forms a conventional paraphase ampli- fier with an 80-volt, 0.25μsec pulse out at the plate and a 15-volt, 0.25 μsec pulse out on the cathode. These two outputs are capacity-coupled to the grids of V3 and V4 respectively.

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(3) Power amplifier V3, V4. Tubes V3 and V4 are connected in push-pull to form a power amplifier. The takeoff point is at the junction of resistors R27 and R22, capacitor C15, and the plate of V4. With no signal in, this point is at zero potential.

(4) Gating circuit. The gating circuit is comprised of tube V5, transformer T1, and tube V6. Tube V6 could be called a gate amplifier itself, but due to the interaction between the various components, V5, V6, and T1 will be called the gating circuit. Capacitor C24 will be referred to as the integrator. The gating circuit and the integrator receive the input error signal and the gate signal to produce a d-c output voltage proportional to the error. This d-c output is effectively controlled by integrator capacitor C24. The output is either positive or negative, depending upon the input signal to the error pulse rectifier. For a positive input input error signal, the resultant output is a positive d-c potential. The opposite is true for a negative input signal.

(5) D-C cathode follower V7A. Cathode follower V7A provides the output current necessary to drive the angle modulator and error meters and to isolate the integrator capacitor C24 from the output.

(6) Gate balance cathode follower V7B. Through the action of V7B, the gate balance circuitry is isolated from the output thus increasing balance stability and ease of adjustment.

(7) Gate balance circuit. The capacitor bank C22, C26, and C27 is used to balance out the residual gate to assure zero output with no input error signal. This is a field adjustment (see TM 9-5000-23).

b. Detail circuit discussion (fig 6-22.1).

(1) Pulse amplifier. Tube V1 and associated circuitry comprise a conventional pulse amplifier. The angle error video pulses appearing at jack J1 are amplified and inverted by V1. The d-c return for the control grid is returned to the midpoint of cathode resistance to obtain proper operating bias for the tube. Resistors R2 and R6 are parasitic suppressors. Capacitors C1 and C2 provide decoupling for the screen grid. The inverse feedback voltage to the cathode from the junction of resistors R30 and R31 is approximately one-tenth the amplitude of voltage appearing at the junction of resistor R27 and capacitor C15. The distorted output of tube V1 is applied through capacitor C3 to the control grid of tube V2. The signal is distorted again in the succeeding circuitry in such a way as to offset this distortion.

(2) Phase inverter and power amplifier. Power amplifier tubes V3 and V4 are connected between -250 volts and +250 volts so that for zero signal

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input, the voltage appearing at the junction of capacitor C15 and resistor R27 is ground. Therefore, negative or positive pulses applied to the power amplifier will appear as negative or positive pulses with respect to ground. Tubes V3 and V4 are connected in a push-pull manner which requires that signals applied to their control grids should be 180° out of phase. This is the function of phase inverter V2 and circuitry. When a negative 15-volt pulse, as shown on figure 6-22.1, is applied from the plate of V1, tube V2 produces an identical 15-volt pulse at its cathode with respect to ground. Also, V2 produces a positive 15-volt pulse on its plate with respect to the potential at the junction of resistors R7 and R8. It can be seen then that the latter pulse is developed across plate load resistor R8. An a-c short is made by capacitors C6 and C10 from the junction of resistors R7 and R8 to the cathode of V3. Therefore, regardless of a change of current through resistor C22, the positive 15-volt pulse from the plate of V2 will appear between the control grid and cathode of V3. At the same instant, the negative pulse from the cathode of V2 is applied to the control grid of tube V4. Since capacitors C13 and C14 provide an a-c ground (decoupler) for the cathode of V4, this pulse appears between the control grid and cathode of V4. Conduction through V4 is decreased, which effectively produces a large cathode resistance for V3. Conduction through V3 is increased. Therefore, more current through an increased cathode resistance produces a large positive pulse at the junction of capacitor C15 and resistor R27. Since the 80-volt pulse is coupled through C6 to the junction of resistors R7 and R8, the pulse appearing at the plate of V2 with respect to ground is 95 volts in amplitude. The power amplifier circuit of V3 and V4 operates far below saturation to reproduce positive and negative video pulses which are symmetrical in shape. In order for V2 to act as a pentode and also have an output at its cathode, capacitor C8 allows the screen grid of V2 to be at the same a-c potential as its cathode. Capacitor C4 is a factory adjustment which provides a small amount of inverse feedback to the plate of V1. This compensates for the tendency of V1 to cause a ringing effect and overshoot of the video pulses. The total resistance of R27 through R31 to ground is 1,106 ohms. The value of resistor R31, therefore, permits approximately one-tenth of the output signal to be applied back to V1. Resistors R9, R14, R16, R23, R19, and R25 are parasitic suppressors. The gain of the amplifier is approximately 20 decibels. The output of the amplifier is coupled through capacitor C15 to the gating circuit.

(3) Receiver gate amplifier. The 15-volt, 0.4µsec receiver gate from the ranging system is applied through jack J2 and coupling capacitor C16 to the control grid of tube V6. This tube is connected as a conventional pulse amplifier whose output is a negative pulse developed across the 4-3 winding of transformer T1. Crystal CR1 is used to prevent overshoot at the end of the gate pulse caused by the partial differentiating action of

the transformer. Any positive signal at terminal 4 with respect to terminal 3 of transformer T1 is shorted through CR1.

(4) Signal gating and d-c cathode follower. Tubes V5 and V7A, transformer T1, and associated circuitry comprise the circuit which generates d-c error signals from the video error signals. Gating is introduced at this stage so that only the video from the target being tracked will affect the d-c output.

(a) When no video error signals are applied to the circuit through capacitor C15, the 15-volt, 0.4μsec receiver gate is applied every 1,000 microseconds to the 3-4 winding of transformer T1. The positive gate on terminal 3 with respect to terminal 4 causes positive signals to appear on terminals 5 and 7 with respect to terminals 6 and 8, respectively. The positive pulses are coupled through capacitors C18 and C21 to the control grids of V5A and V5B, respectively. Since the cathodes of the two triodes are so connected to transformer T1, the positive gate appears between the grid and cathode of each tube. Since the assumption was made that no video signals are applied through capacitor C15, there will be no tube conduction in V5A and V5B because tube halves are held cut off by the negative voltage on the grids in respect to the cathodes. With no input signal coupled through capacitor C15, there is no potential difference between the plate and cathodes of V5A and V5B. However, these tubes are not restricted from conduction should a potential difference between cathodes and plates exist. During the gate period a small amount of grid current is drawn through resistors R36 and R37. As a result capacitors C18 and C21 charge negatively. When the gate is removed from transformer T1, these capacitors discharge slowly through their respective resistors. Current flow through resistors R36 and R37 is such that the grids of V5A and V5B become sufficiently negative with respect to their cathodes to bias the tubes below cutoff. Therefore the tubes are not permitted to conduct even though a voltage might appear between plate and cathode. When the next gating pulse is applied from transformer T1, the amplitude of the gate overrides the bias voltage on the grids and allows grid conduction. The pulse also causes the two capacitors to charge again, thus replacing the negative charge which leaked off between gate pulses. This is the gating action of the circuit. Only those video pulses applied during the gate will be allowed to cause plate conduction in V5A and V5B.

(b) When, during the application of a gating pulse, a positive 80-volt pulse is coupled through capacitor C15, this signal is then present on the cathode of V5A and plate of V5B. Through this action V5A will remain cut off. V5B, however, will conduct heavily. Therefore the error pulse immediately charges capacitors C15 and C24. The electron flow in this

charging process is from the top side of capacitor C24, through resistor R38, V5B, resistor R45, capacitor C15, resistors R27, R28, R29, R30, and R31 to ground, then from ground to the ground side of capacitor C24. The capacitance of C24 is a fraction of capacitor C15. For all practical purposes, then, the 80-volt pulse appears across capacitor C24. When the gating pulse and video pulses are removed, the gating action of the circuit prevents conduction in V5A and V5B. The charge on capacitor C24 remains the same until the next gating pulse. If the next video pulse (occurring with the gate pulse) is a positive 60 volts, the cathode of V5A will be a minus 20 volts in respect to the plate of V5A, causing V5A to conduct. This causes capacitor C24 to charge to the value of the new input signal of 60 volts. This process is repeated until the antenna is pointed on-target, at which time no video error signal is applied to the gating circuit. At this time capacitor C24 has no charge on it.

(c) The circuit action above can be applied in the case of negative video error pulses, in which case capacitor C24 holds a negative charge with respect to ground. Winding 1-2 of transformer T1 along with cathode follower V7A and GATE BAL variable capacitor C22 are integrated into the unit to compensate for residual gate signals. This will assure that capacitor C24 will be able to have a zero charge when there is no video error signal input to the gate circuit. The d-c error signal is applied from capacitor C24 through parasitic resistor R39 to the grid of V7B. Loss of signal strength from the input to the gating circuit to the output of cathode follower V7B is approximately 6 decibels. Tube V7B is a conventional cathode follower with its cathode circuit returned to a regulated -250 volts. This makes possible the output of negative as well as positive signals. The output is applied through terminal 12 of plug P1 to a GS-15527 angle modulator.

(5) Zero adjust and monitoring. Refer to figure 6-23.1. A portion of the d-c output of this unit is also coupled directly through pin 15 of connector P2 to the AZ ANGLE ERROR meter on the GS-15597 track test panel. Associated with the AZ ANGLE ERROR meter is a ZERO ADJUST potentiometer R5 in the test panel which is used to adjust the d-c potential on capacitor C24. By adjustment of the ZERO ADJUST potentiometer in the test panel, the quiescent charge across capacitor C24 in the error pulse rectifier is changed to the proper value to obtain zero d-c potential out of the unit indicated by the AZ ANGLE ERROR meter.

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